THE ELECTRONICS MAGAZINE WITH THE PRACTICAL APPROACH

UK £1·60 IR £2·35 (incl. VAT) February 1989

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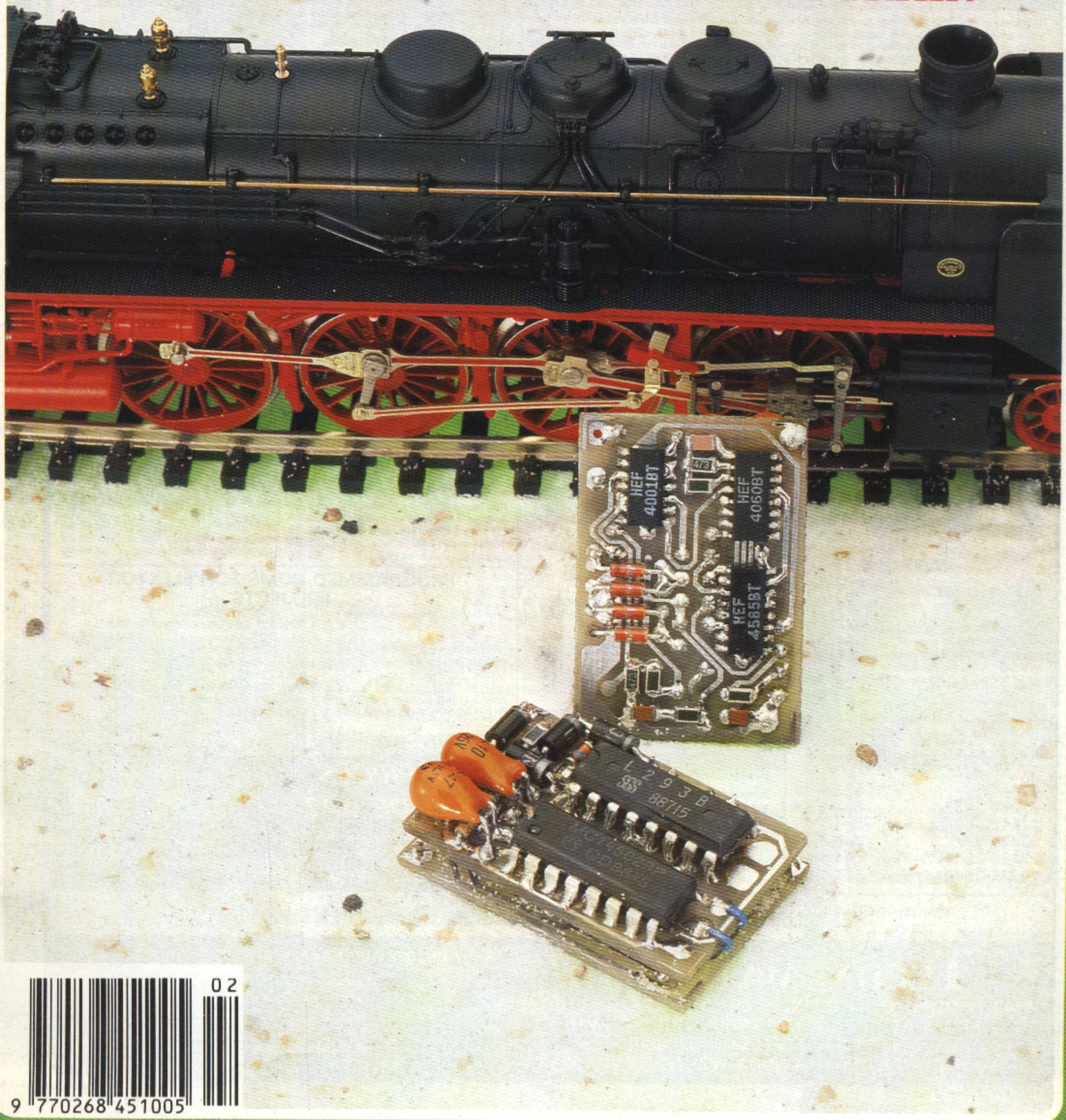
Elektor Electronics

Morse code generator
Videocards in PCs
Counter without counter
VHF receiver
Touch key organ
Wideband RF amplifiers
Dark-room timer
Tweeter protector

NEW SERIES: THE DIGITAL MODEL TRAIN



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Elektor Electronics

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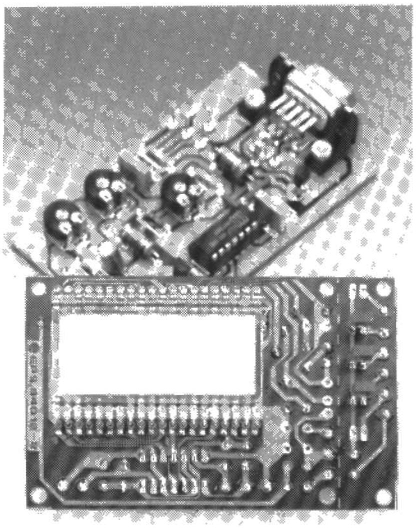

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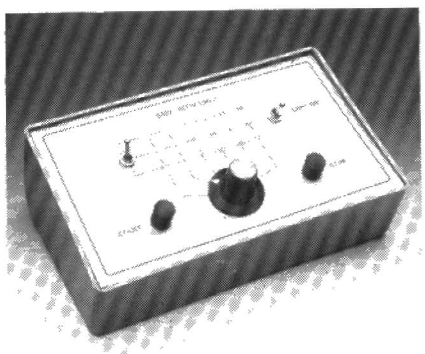

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- Wide-band amplifier*
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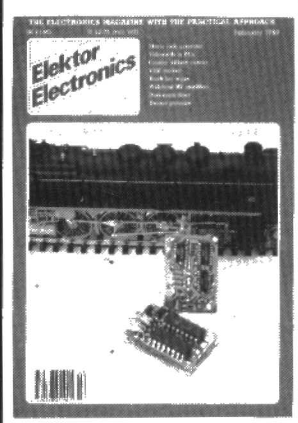
- Low-battery indicator
- Dealing with e.m.i.
- Counter without counter*

* We regret that owing to circumstances beyond our control this project could not be included in this issue.



Front cover

This month's front cover illustrates the beginning of a new series of articles describing a number of model railway units based on new technology. The series culminates in a fully electronic model railway.

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Distribution:
SEYMOUR
1270 London Road
LONDON SW16 4DH.
Typeset & composed in the Netherlands by GBS, Beek (L). Printed in the Netherlands by NDB, Zoeterwoude.

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MEMBER OF THE AUDIT BUREAU OF CIRCULATIONS

ELECTRONICS AND PROTECTIONISM

Protective tariffs, (call them levies, moderations, limitations, quotas, subsidies, the name does not matter: they all mean non-productive penalty), so eagerly sought by some and so detested by many, are bad for the consumer because he has to foot the bill. In any case, as history has proved time and again, such protectionism offers no long-term solution to what is, basically, an inefficient industry or economy. On the contrary, it causes an industry or economy to go soft, thereby exacerbating the real problem. What we need and want is a Europe that is based on strong industries and economies, that feeds on competition, and that can and does hold its own in the world without protective measures.

We hear much of unfair competition, of goods made outside the European community and sold in it at unrealistic prices. But what is unrealistic? And who decides that? Under community law, the EC Trade Commission decides what is an unrealistic price. But, as illustrated by a recent BBC TV programme, the trade commission has a strange — though legal — way of arriving at the cost of, say, a video recorder or printer. They deduct from the price charged in the EEC for that equipment all the direct and indirect operating expenses, but — and here is the crux of the matter — from the price charged in the country of origin only the direct operating costs. The difference between the two 'basic costs' is, in the commission's parlance, the 'dumping margin'. Note that this margin is equal to the supplier's indirect operating costs, which include, for instance, repayment of investment expenditure.

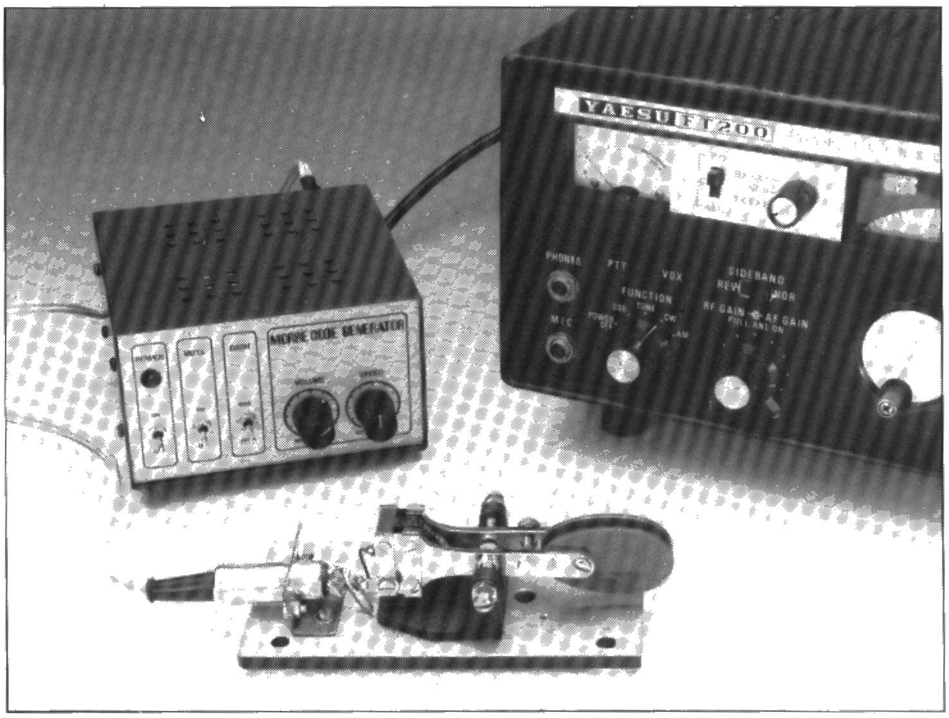
To most people, dumping is selling at a loss to gain market share. In all cases we know of, the supplier has stated, and is willing to prove, that he is making a reasonable profit. None the less, and entirely within the laws of the community, the commission is determined that a levy be paid on 'dumped' equipment. This levy is not going to be taken from the supplier's profit, but is going to have to be paid by us, the consumers. So, prices of a miscellany of electronic goods are, or will be, raised artificially. In other words, we, the consumers, pay, by order of the EEC, a penalty on equipment that is considered too cheap compared with similar equipment made in Europe.

One might be forgiven for thinking that these penalties are levied only on goods originating in the Far East. However, it is also happening to equipment manufactured in Britain, no, not Nissan cars, but video recorders made by Amstrad in their Essex factory. It is a fact that a large proportion of components in this machine are of Far Eastern origin, but that is the case in much electronic equipment. In our modern world of high technology and cross-fertilized, trans-continental industries, it is next to impossible in many cases to trace the origin of an electronic device.

There is not much sympathy for the consumer in Brussels. There is a body of opinion that the inhabitants of all EEC states live in an electronic paradise with low prices and abundant choice. Those who are of that opinion should go and shop around in electronics stores outside the EEC. It is illuminating that last November's issue of *What to buy for Business* claimed that British office equipment buyers are being badly treated by having to pay more for their Japanese printers than they need "simply to protect the profit margins of a handful of European producers". The same magazine also decries the hypocrisy of some EEC suppliers who often sell Far Eastern equipment under their own brand names.

Fortunately, the commission's anti-dumping policy has been referred to the United Nations' General Agreement on Tariffs and Trade (GATT), who have subsequently opened an inquiry into the complaints. For all our sakes, let us hope that common (economic) sense will prevail and that the EEC Trade Commission will be persuaded to move away from the dangerous path of protectionism.

MORSE CODE GENERATOR



Ideal for both the morse trainee and the experienced operator, this low-cost, versatile, generator with relay output provides automatic timing, at a user-defined speed, of audible dots and dashes.

by D. McBright

In this circuit, four 74HC series ICs are used to produce either a string of dots, or a string of dashes. The dash has three times the length of a dot. Spacing between characters is equal to one dot. Characters are selected by shorting either of two contacts of a paddle-type key to ground. The generator allows the string of 'automatic' dashes to be replaced by dashes whose length and frequency are controlled by the operator, whilst the string of dots remains unchanged. The frequency of the dots ranges between about 130 and 910 per minute. A variable-level sidetone is available at a fixed frequency to enable the operator to hear the characters he is sending.

Output from the generator is a normally-open relay contact for connecting to the CW transmitter. A switch is included to disable the morse relay and so reduce the current consumption while practising. The morse generator is powered by a 9 V battery or by a mains adaptor with an output between 8 and 15 VDC. An internal voltage regulator supplies 5.6 V for the integrated circuits.

Circuit description

The circuit diagram is given in Fig. 1. When counting, the first divider, IC_1, divides the clock signal supplied by N_1 by 16. The oscillator operates at 128 times the dot frequency, and uses the hysteresis of a CMOS NAND gate, N_1, to give a charge-discharge cycle for R-C network (P_2+R_8)-C_8. The second divider in the circuit, IC_2, uses its first 2 stages to divide by 4; the third stage to produce dots or the first third of dashes; the fourth stage — in conjunction with the third — to produce dashes. Gates N_4 and N_3 are the respective inverters, and N_6 inhibits the final two-thirds of the dashes when dots are required. Between counts, divider 1 is reset to 15, and divider 2 is reset to 3 so that, when the appropriate contact is shorted, a character is started on the next positive edge from the oscillator. The maximum delay between a contact being actuated and the start of a character is approximately 3.5 ms at the slowest dot speed.

When either of the 2 key inputs of 3-input NAND gate N_5 is connected to ground, the load inputs of the dividers are taken logic high, and counting is started. The third input of N_5 is used in conjunction with feed-back diodes D_1 and D_2 to ensure that any character is completed if a key contact is broken early (this does not apply to dashes in the manual mode). However, a full-length dash will only be obtained if the contact is released after the first third of the character is completed, otherwise a dot of the correct length is produced.

Two of the three inputs of NAND gate N_7 mix the pulses from the dot/dash inverters, N_3 and N_4, whilst the third is keyed only in the 'manual dash' mode to make non-automatic dashes. Automatic or manual operation is selected by toggle switch S_1. The output of N_7 controls the sidetone oscillator set up around N_2, and the relay driver, T_1. Optimum sound output from the passive piezo-ceramic buzzer Type PB2720 from Toko is stated to be at a frequency between 3 and 3.5 kHz. Reasonable sound levels are, however, obtained at lower frequencies also. The sidetone oscillator frequency can be adjusted to individual taste by altering the value of R_7 between 68 kΩ (min.) and 220 kΩ (max.)

A stabilized +5.6 V is provided by a 5 V/100 mA regulator, IC_5, whose output voltage is raised by 0.6 V with the aid of a conducting silicon diode, D_3, connected between the common terminal and ground. As already noted, the circuit can be powered from a battery of a mains adaptor with DC output. In stand-by mode, the generator draws 6-7 mA from a 9 V supply; with the relay muted and characters selected, current consumption rises to about 10 mA. Total current consumption with the relay actuated depends mainly on the coil resistance.

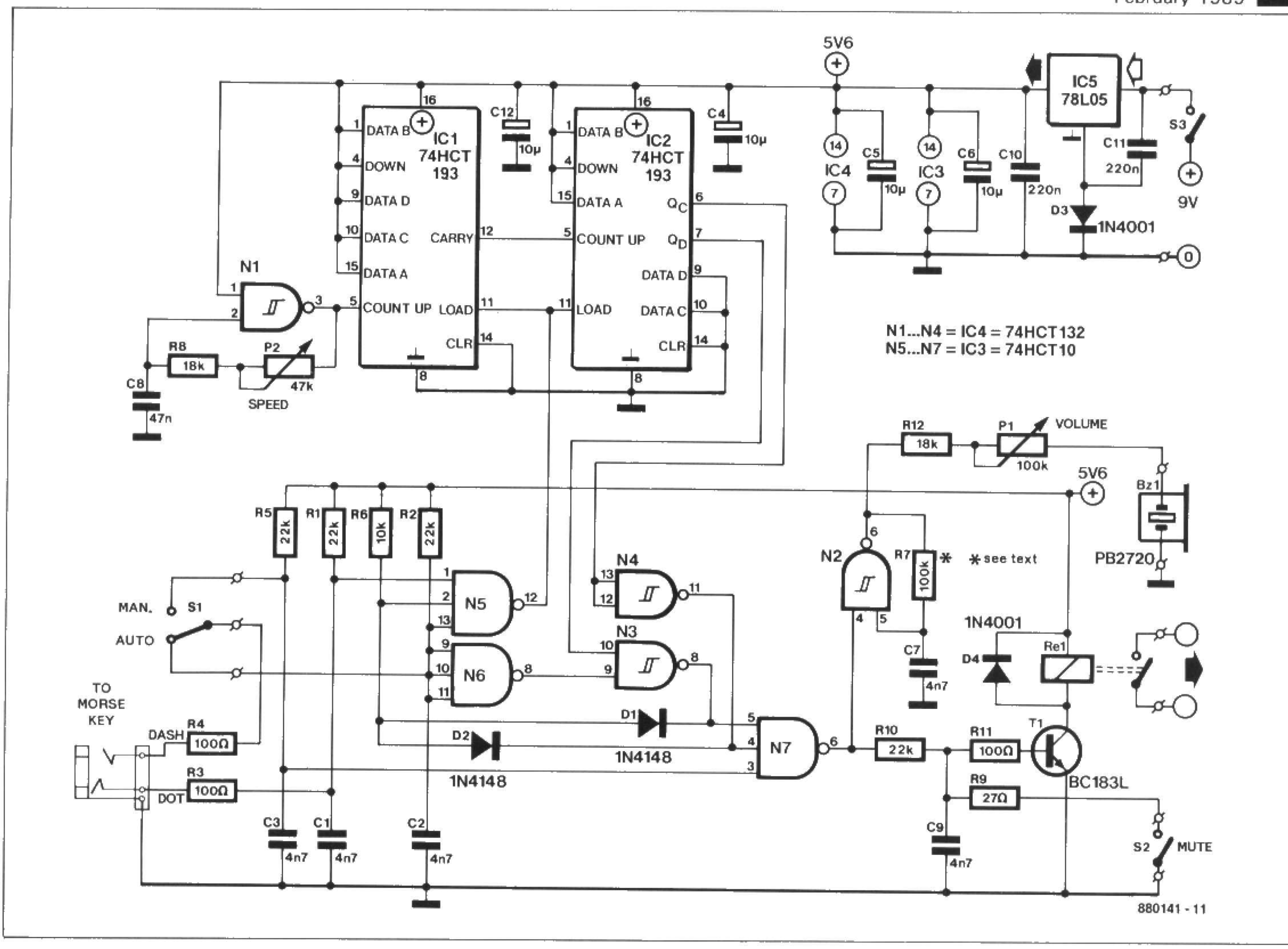


Fig. 1. Circuit diagram of the morse code generator with relay and sounder output.

Construction

The generator is conveniently built on a small piece of veroboard or other prototyping board. Construction and wiring are straightforward. All ICs are fitted in sockets, and solder terminals are provided for the wires to the external controls. Whatever type of relay is used, a diode to suppress back-EMF *must* be provided as shown in the circuit diagram (this diode is integral to most, but not all, types of DIL reed relay operating from 5 V). The coil resistance of Re_1 should not be lower than about 500 Ω.

The accompanying photograph shows the completed morse code generator connected to an all-mode SW transceiver Type FT-200 from Yaesu. The generator is fitted in a small metal enclosure, with all controls mounted onto the front panel and connected to the board via wires and solder terminals. The piezo buzzer is glued in place behind the vent holes at the inside of the top lid of the enclosure. The paddle-type key is connected to the generator via a short length of stereo screened wire and a 3.5 mm stereo headphone plug (fitted on to the rear panel) and mating socket. The screening of the wire is connected to the centre contact of the paddle.

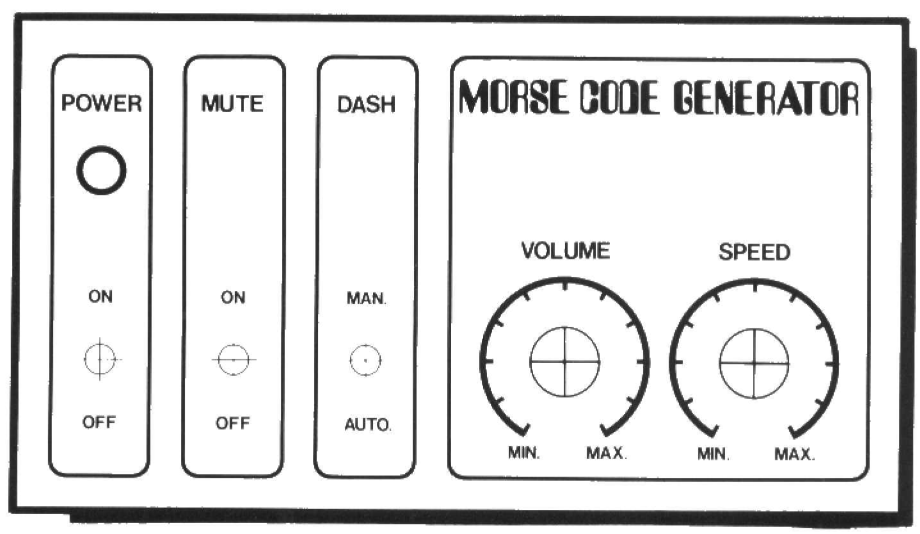


Fig. 2. Suggested front-panel layout.

MOSFET HI-FI POWER AMPLIFIER

A quality 160-watt hi-fi output amplifier based on the Siemens BUZ series MOSFETs.

Until not so long ago, the BUZ series of MOSFETs from Siemens were hard to come by and very expensive. That was a pity, because these devices offer a very good specification. Fortunately, the situation has improved considerably, although the transistors are still only available as n-p-n types. However, n-p-n types can be used just as well as complementary pairs as the present circuit proves.

A power amplifier, whether it uses bipolar devices or MOSFETs, needs a drive circuit. When MOSFETs are used, that circuit can be kept pretty straightforward. This means that any modifications in respect of power handling, bandwidth and distortion may be brought in fairly easily.

The device chosen for the present circuit is recommended by Siemens for use as a power opamps in control engineering, which indicates that it is a very stable component. None the less, to prevent any mishaps, the amplifier is provided with protection circuits against short-circuits and overheating.

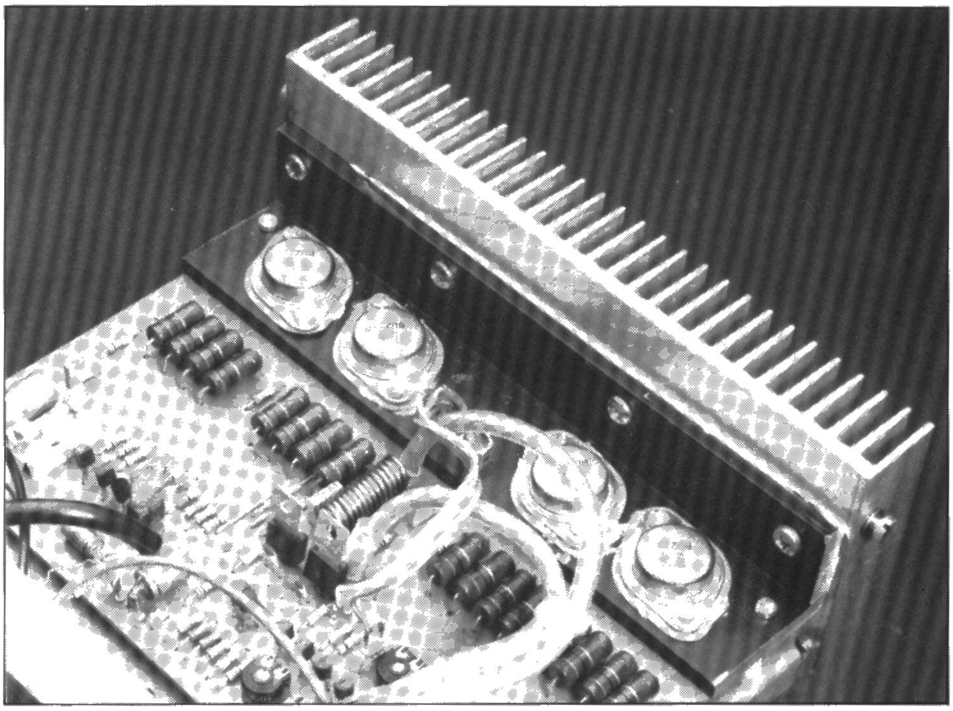


Fig. 1. The completed MOSFET power amplifier.

The circuit

The circuit diagram in Fig. 2 shows the highest-power version of the amplifier: this delivers 160 W into 4 ohms. Modifications to reduce the power ouput will be discussed later in the article.

The circuit is based on the two series-connected MOSFETs, T_{15} and T_{16}, being driven in anti-phase by a differential amplifier. Since the input resistance of MOSFETs is of the order of 10^9 ohms, the drive power needs to be only very small. The MOSFETs are thus voltage-driven.

The drive circuit consists essentially of T_1-T_2 and T_{12}-T_{13}. Negative d.c. feedback from the output amplifier is provided by R_{22} and negative a.c. feedback by R_{23}-C_3. The a.c. voltage gain is about 30 dB. The lower cut-off frequency depends on the values of C_1 and C_3.

The operating point of the first differential amplifier, T_1-T_2, is set by the current flowing through T_3. The collector current of T_5 determines the reference current for current mirror T_3-T_4. To ensure that the reference current is stable, the base voltage of T_5 is stabilized by diodes D_4-D_5.

The output of T_1-T_2 drives a second differential amplifier, T_{12}-T_{13}, whose collector currents generate the gate potential for the output transistors. The level of that potential is determined by the operating point of T_{12}-T_{13}. Current mirror T_9-T_{10} and diodes D_2-D_3 have the same function as T_3-T_4 and D_4-D_5 in the first differential amplifier. The magnitude of the reference current depends on the collector current of T_{10}, which in turn is set by P_2 in the emitter circuit of T_{11}. This arrangement sets the quiescent (bias) current in the absence of an input signal.

Stabilization of quiescent current

The MOSFETs have a positive temperature coefficient when their drain current is small, so that the quiescent (bias) current is only kept stable by appropriate compensation. This is provided by R_{17}

D.C. operating voltage	
(P_{out} = max)	≥ ±46 V
(P_{out} = 0)	≤ ±55 V
Current drawn	
(P_{out} = max)	3 A
(P_{out} = 0)	≥0.2 A
(Output short-circuited)	≤1.5 A
Max. power output	
(f = 1 kHz; R_L = 4 ohms)	160 W
Music power output	
(R_l = 4 ohms)	≤240 W
Distortion (20 Hz-20 kHz)	≤0.05%
Intermodulation	
(250 Hz; 8 kHz; 4:1)	≤0.07%
Input resistance	≤33 k
Voltage amplification	31 dB
Frequency response (−3 dB)	≥2 Hz-≤250 kHz
(R_L = 4 ohms; P_{out} = 15 W)	
Power bandwidth	
(THD = 0.5%; P_{out} = 80 W)	≤5 Hz-≥70 kHz
Damping factor	
(R_L = 4 ohms; f = 40 Hz)	≥200
Signal-to-noise ratio (unweighted)	
(P_{out} = 50 mW)	≥73 dB
(P_{out} = max)	≥108 dB
Output impedance	4 Ω

Table 1. Technical specification of the MOSFET amplifier.

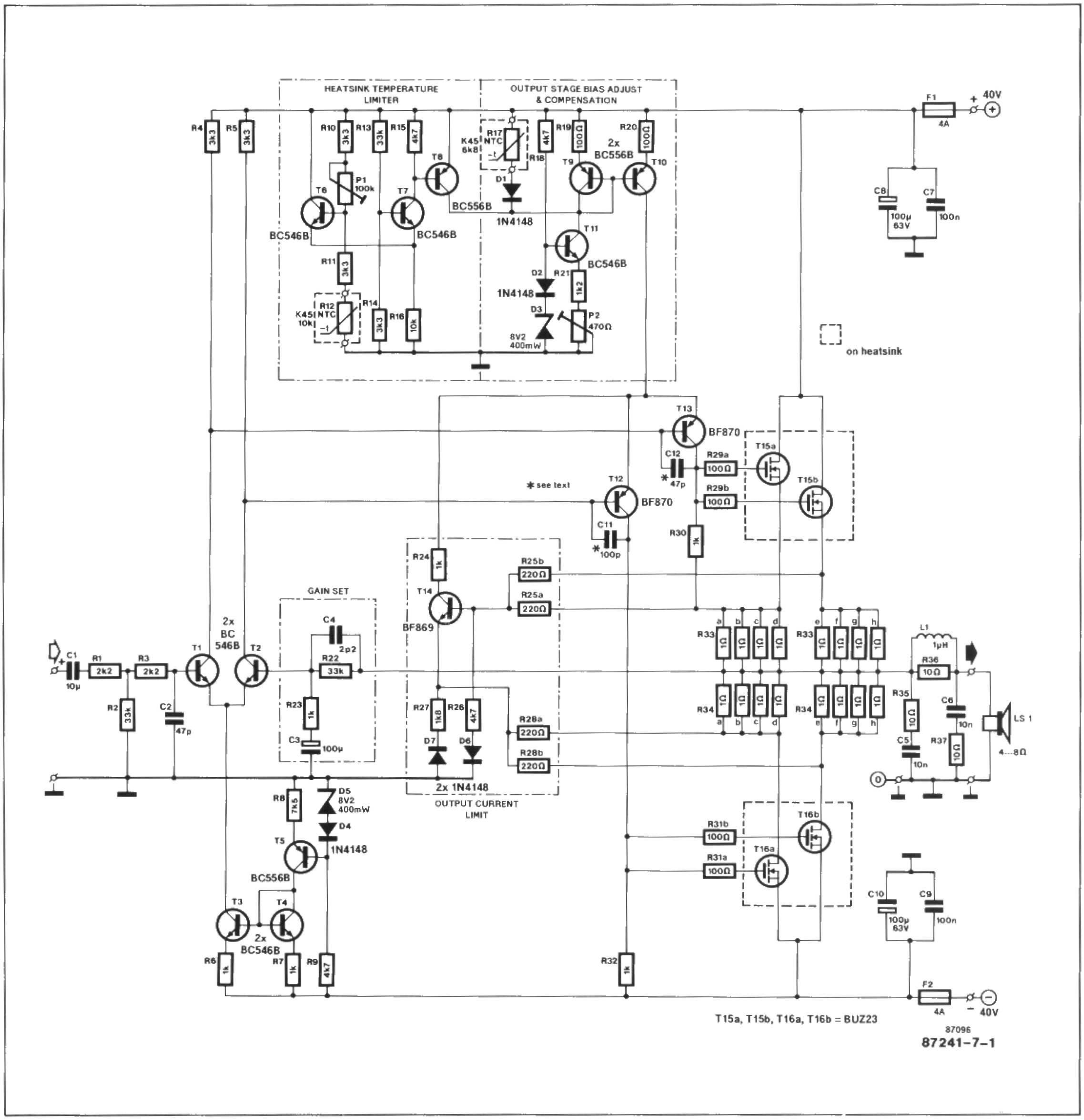


Fig. 2. Circuit diagram of the 160-watt version of the MOSFET power amplifier. Changes for lower-power versions are given in Table 2.

across current mirror T_9-T_{10}, which has a negative temperature coefficient. When this resistor heats up, it draws a slightly larger portion of the reference current through T_9. This causes a reduction in the collector current of T_{10} and this, in turn, causes a decrease in the gate-source voltage of the MOSFETs, which effectively compensates the increase caused by the PTC of the MOSFETs. The thermal time constant, which is dependent on the thermal resistance of the heat sinks, determines the time it takes for stabilization to be effected. The quiescent (bias) current set by P_2 is stable within ±30%.

Overheating protection

The MOSFETs are protected against overheating by thermistor R_{12} in the base circuit of T_6. When a certain temperature is reached, the potential across the thermistor causes T_7 to switch on. When that happens, T_8 draws the larger part of the reference current through T_9-T_{11}, which effectively limits the output power of the MOSFETs.

The temperature threshold is set by P_1 and is equivalent to a heat sink temperature of ≤72.5 °C. This assumes a thermal resistance of 0.5 K W and an ambient temperature of 25 °C.

Short-circuit protection

If the output is short-circuited in the presence of an input signal, the reduction in voltage across resistors R_{33} and R_{34} causes T_{14} to be switched on. This results in a decrease of the current through T_9-T_{10} and, consequently, of the collector currents of T_{12} and T_{13}. The dynamic range of the MOSFETs is then severely restricted, so that the power dissipation is kept low.

Since the permissible drain current is dependent on the drain-source voltage, more information is needed for the correct setting of the current limiting. This

Parts list

Resistors:
R_1;R_3 = 2k2
R_2;R_{13};R_{22} = 33 k
R_4;R_5;R_{10};R_{11};R_{14} = 3k3
R_6;R_7;R_{23};R_{24};R_{30};R_{32} = 1 k
R_8 = 7k5
R_9;R_{15};R_{18};R_{26} = 4k7
R_{12} = 10 k(NTC)*
R_{16} = 10 k
R_{17} = 6k8 (NTC)*
R_{19};R_{20};R_{29a};R_{29b};R_{31a};R_{31b} = 100R
R_{22} = 1k2
R_{25a};R_{25b} = 330R
R_{27} = 1k8
R_{28a};R_{28b} = 220R
$R_{33a\text{-}h}$;$R_{34a\text{-}h}$ = 1R; 1 W
R_{35};R_{37} = 10R
R_{36} = 10R; 1 W
P_1 = 100 k preset
P_2 = 470R preset
* Siemens Type K45 or equivalent

Capacitors:
C_1 = 10μF (MKT)
C_2 = 47pF
C_3 = 100μF; 16 V
C_4 = 2p2
C_5;C_6 = 10nF
C_7;C_9 = 100nF
C_8;C_{10} = 100μF; 100 V
C_{11} = 100pF; 100 V
C_{12} = 47pF; 100 V

Semiconductors:
D_1;D_2;D_4;D_6;D_7 = 1N4148
D_3;D_5 = 8V2; 0.4 W (zener)
T_1;T_2;T_3;T_4;T_6;T_7;T_{11} = BC546B
T_5;T_8;T_9;T_{10} = BC556B
T_{12};T_{13} = BF870
T_{14} = BF869
T_{15a};T_{15b};T_{16a};T_{16b} = BUZ23

Miscellaneous:
L_1 = 15 turns enamelled copper wire 0.2 mm dia. wound on R36
F_1;F_2 = miniature fuse, 4 A on PCB holder
Right-angle aluminium extrusion as shown in Fig. 4.

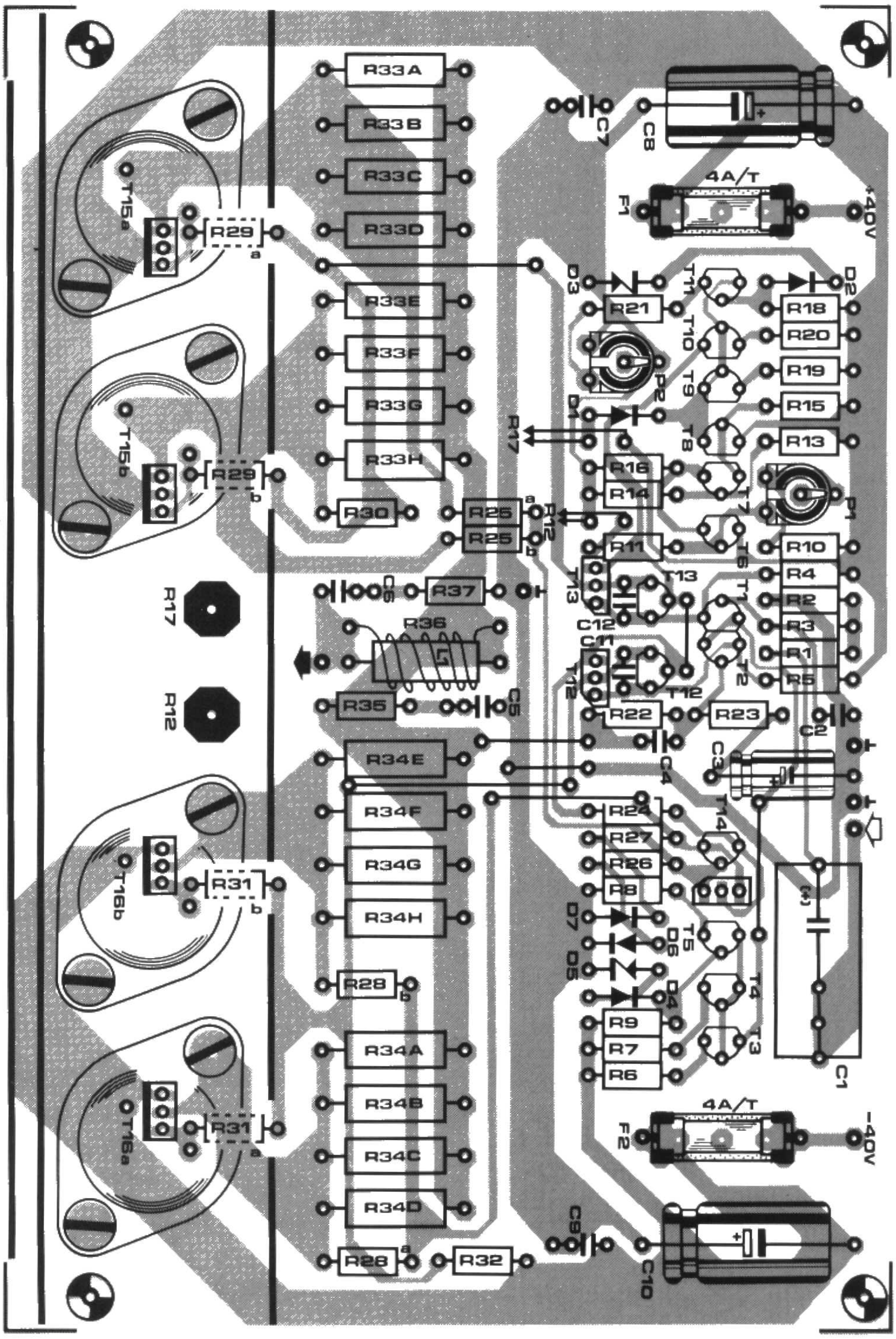


Fig. 3. The printed-circuit board for the MOSFET power amplifier. The associated parts list is intended for the 160-watt version. Changes for lower-power versions are given in Table 2.

information is provided by the voltage drop across resistors R_{26} and R_{27} (positive and negative output signals respectively). If the load is >4 ohms, the base-emitter voltage of T_{14} is reduced to a value that results in the short-circuit current being limited to 3.3 A.

Construction

The amplifier is best build on the PCB shown in Fig. 3. However, before construction is started, it has to be decided which version is wanted. Fig. 2 and the parts list of Fig. 3 are for the 160-watt version. Changes for the 60 W, 80 W, and 120 W versions are shown in Table 2.

As shown in Fig. 4, the MOSFETs and NTCs are mounted on a right-angled. The pin connections are shown in Fig. 5. The NTCs are screwed direct into M3-size, tapped (tapping drill = 2.5 mm), holes: use plenty of heat conducting paste.

Resistor R_{29} and R_{31} are soldered direct to the gates of the MOSFETs at the track side of the board.

Inductor L_1 is wound on R_{36}: its well-insulated, pre-tinned terminals are soldered to the holes adjacent to those for R_{36}.

Capacitor C_1 may be an electrolytic type, but an MKT type is preferable.

The faces of T_1 and T_2 should be glued together to ensure that their body temperature remains equal.

Do not forget the wire bridges.

The power supply for the 160-watt version is shown in Fig. 6: changes for the other versions are shown in Table 2. An artsist's impression of its construction is shown in Fig. 7.

Once the power unit has been built, the open-circuit operating voltages may be measured. The d.c. voltages should be not greater than ± 55 V, otherwise there is a danger that the MOSFETs will give up the ghost on first power-on. If suitable loads are available, it is, of course, preferable that the supply is tested under load conditions.

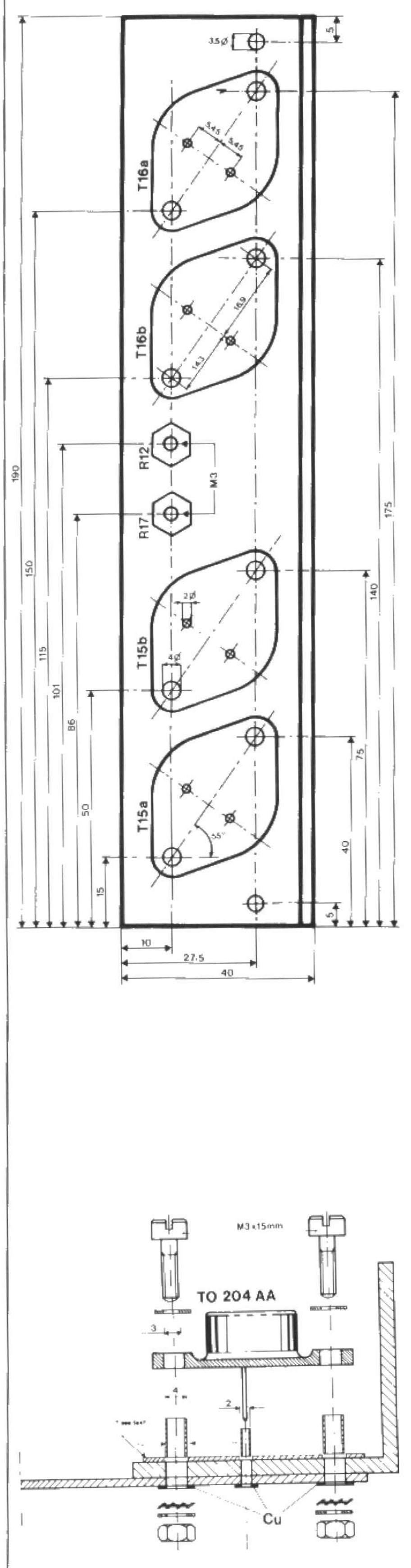


Fig. 4. The right-angle aluminium bracket on to which the MOSFETs and NTCs are mounted. The bracket itself is fitted on to the printed-circuit board.

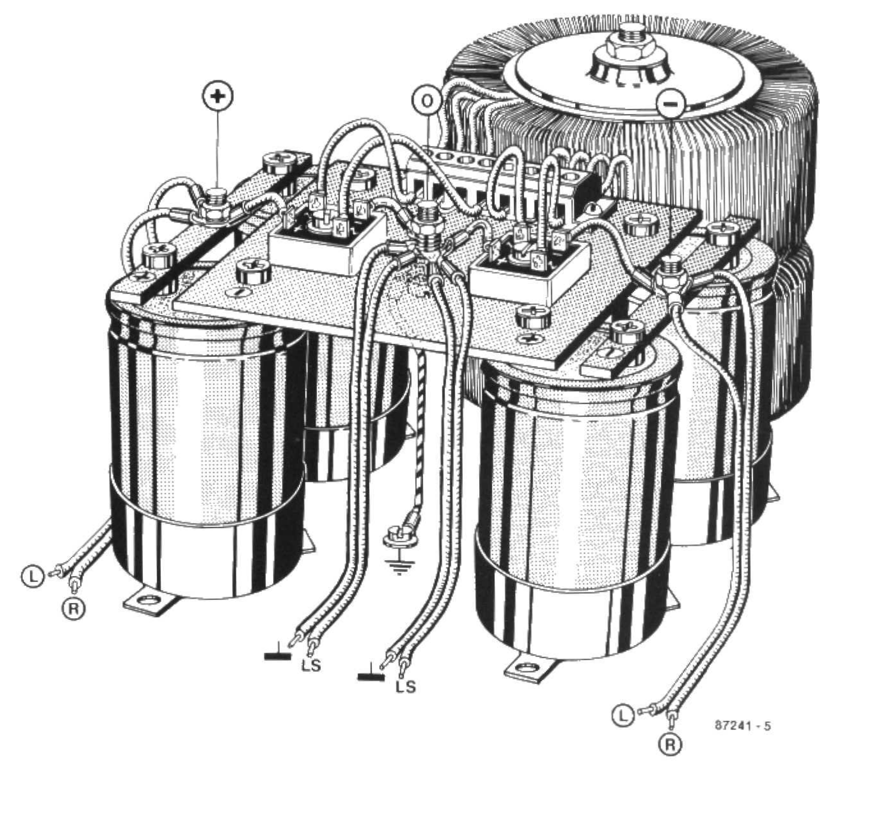


Fig. 5. Artist's impression of the construction of the power unit.

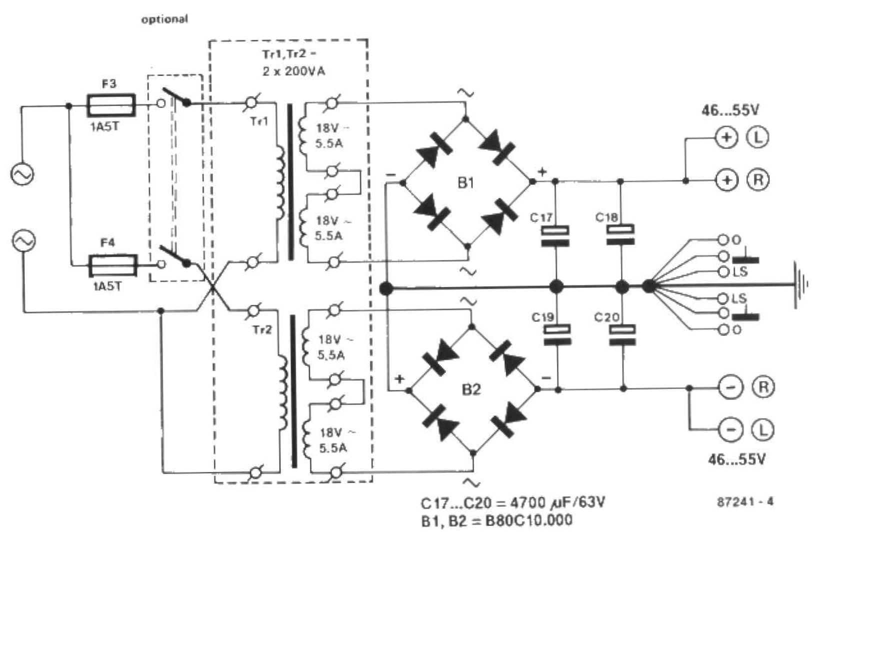


Fig. 6. Circuit diagram of the power unit for the MOSFET amplifier.

Parts list

B1;B2 = bridge rectifier 100 V; 25 A

C17;C18;C19;C20 = 4700–10,000 µF; 63 V

F3;F4 = miniature fuse 1.5 A

Tr1;Tr2 = mains transformer with 2 × 18 V; 5.5 A (200 VA) secondary

When the power supply is found OK, the aluminium MOSFET assembly is screwed on to a suitable heat sink. Fig. 8 gives an impression of the size of the heat sinks and of the complete assembly of a stereo version of the amplifier. For clarity, only the position of the components of the power supply is shown.
The areas where the heat sink and the aluminium MOSFET assembly (and, possibly, the rear panel of the amplifier enclosure) meet should be given a good coating of heat conducting paste. Each of the two assemblies should be screwed to the associated heat sink with at least six M4 (4 mm) size screws.
The wiring should follow the guide lines in Fig. 8 faithfully. It is best to start with the supply lines (heavy gauge wire). Next, make the earth connections (star-shaped) from the power unit earth to the PCBs and the output earth. Subsequently, make the connections between PCBs and loudspeaker terminals and those between the input sockets and the PCBs. The input earth needs to be connected only to the earth terminal on the PCB - no more!

Calibration and testing

Instead of fuses F_1 and F_2, connect 10-ohm, 0.25 W, resistors in their position on the PCB. Preset P_2 must be set fully anticlockwise, while P_1 is set to the centre of its travel. The loudspeaker terminals remain open, and the input must be short-circuited.
Switch on the mains. If there are any short-circuits in the amplifier, the 10-ohm resistors will go up in smoke! If that happens, switch off immediately, find the fault, replace the resistors, and switch on again.
When all appears correct, connect a voltmeter (3 V or 6 V d.c. range) across one of the 10-ohm resistors. There should be no voltage across it. If there is, P_2 is not turned fully anticlockwise. The voltage should rise when P_2 is slowly turned clockwise. Set P_2 for a voltage of 2 V: the current is then 200 mA, i.e.: 100 mA per MOSFET.
Switch off, and replace the 10-ohm resistor by the fuses. Switch on again, and measure the voltage between earth and amplifier output: this should be not greater than ± 20 mV. The amplifier is then ready for operation.
A final point. As already stated, the switching threshold of the overheating protection circuit must be set for about 72.5 °C. This can be ascertained by heating the heat sink with, e.g., a hair dryer and measuring its temperature. However, this is not strictly necessary: P_1 may be left set at the centre of its travel. Its position should only be adjusted if the amplifier switches off too often. None the less, its position should never be far from the mid position. ◀

Power transistors:		2 × BUZ 20	2 × BUZ 23	4 × BUZ 20	Unit
DC operating voltage					
(P_{out} = max)	≥	± 33	± 36	± 40	V
(P_{out} = 0)	≤	± 38	± 42	± 50	V
Current drawn					
(P_{out} = max)		0,1	0,1	0,2	A
(P_{out} = 0)	≥	1,7	2	2,3	A
(output short-circuited)	≤	1	1	1,8	A
Power output (max)					
(f = 1 kHz; R_L = 4Ω)		60	80	120	W

Transistors required

Transistors	60 W	80 W	120 W
T_1, T_2	BC 414 C	BC 414 C	BC 546 B
T_3, T_4	BC 237 B	BC 237 B	BC 546 B
T_5	BC 307 B	BC 307 B	BC 556 B
T_6, T_7	BC 237 B	BC 237 B	BC 546 B
T_8, T_9, T_{10}	BC 307 B	BC 307 B	BC 307 B
T_{11}	BC 237 B	BC 237 B	BC 546 B
T_{12}, T_{13}	BC 556 B	BC 556 B	BF 870
T_{14}	BC 546 B	BC 546 B	BF 869
T_{15a}, T_{16a}	BUZ 20	BUZ 23	BUZ 20
T_{15b}, T_{16b}	–	–	BUZ 20

Resistors for short-circuit protection	R25a,b	R28a,b	R26	R27	Unit
60/80 W	330	120	2,7 k*)	1 k*)	Ω
120/160 W	330	220	4.7 k*)	1,8 k*)	Ω

*) The onset of the short-circuit protection is determined by these values and must be adapted for each and every individual amplifier.

Table 2. Changes and variations for lower-power versions of the MOSFET amplifier.

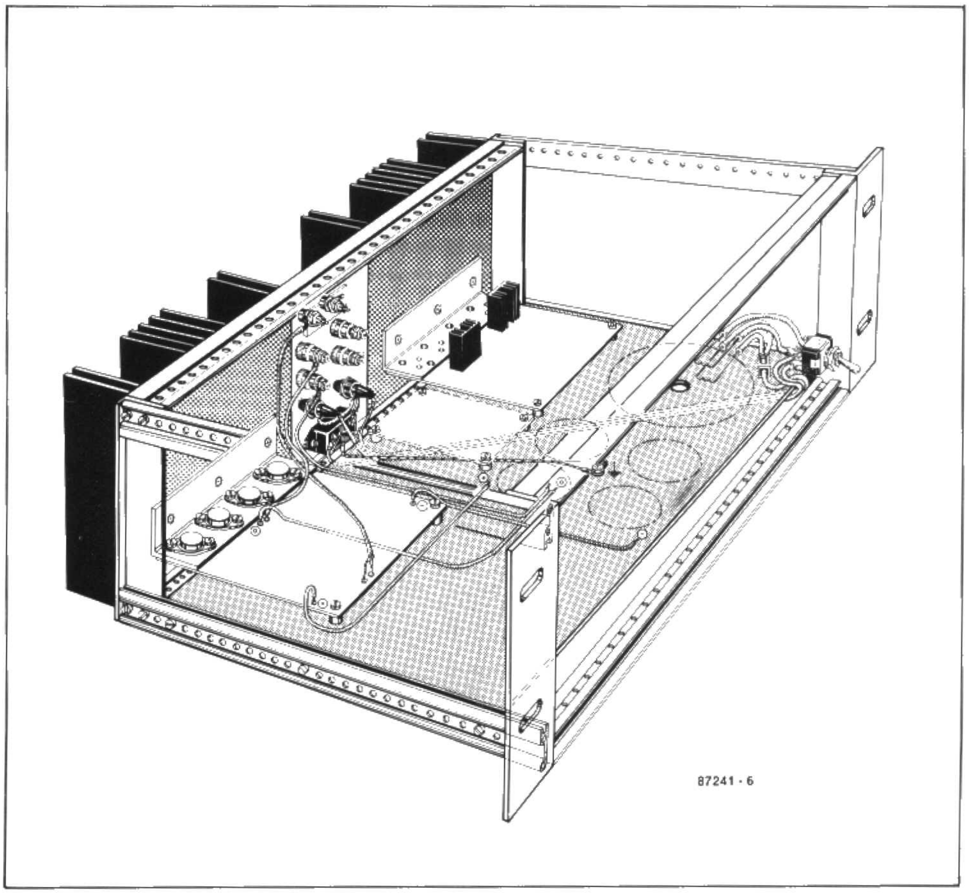
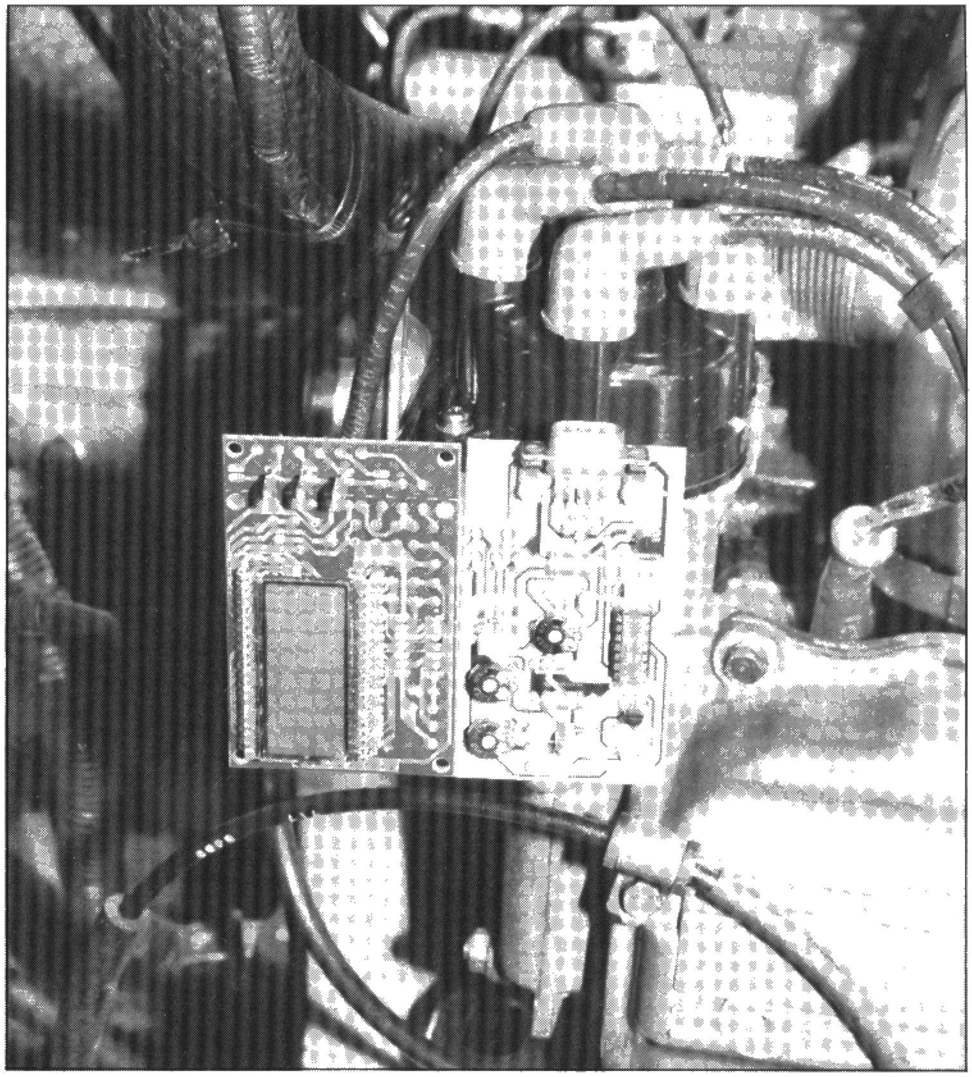


Fig. 7. Artist's impression of the assembly of a stereo version of the MOSFET power amplifier. It also gives an idea of the size of the heat sinks.

CAR SERVICE MODULE

A compact unit that measures speed of a petrol engine in revolutions per minute, and the dwell angle of the ignition.

by. A. Rigby



The car service module is composed of two units: a circuit for measuring dwell angle and engine speed on one printed circuit board, and an associated liquid crystal display (LCD) read-out on another board. The units are connected by a cable terminated in 9-pin D-type connectors. The compact LCD readout is purposely kept separate to enable it to be used in other applications also.

Electronics and the petrol engine

Engine speed and the ignition dwell angle are both physical quantities which are to be converted to a voltage that can be shown on a display. Figure 1a shows the basic elements of an ignition system in a petrol engine. The primary of the ignition coil is connected between the positive pole of the car battery and the contact breaker, which is shunted by a capacitor and indirectly operated by the camshaft. When the camshaft revolves, the contact breaker opens periodically. A magnetic field is built up in the ignition coil when the contact breaker is closed. When the contact opens very briefly as it is pushed open by the rotation of the camshaft, the magnetic field causes an electrical pulse because of resonance of the tuned circuit formed by the ignition coil and the capacitor. The alternating voltage is boosted to 15,000 to 30,000 volts by the high-impedance secondary winding of the coil. The high voltage is then directed, via the distributor, to one of the 4 spark plugs (it is assumed here that the service module is used for 4-cylinder cars). Obviously, the spark rate depends on the speed at which the engine runs.

The dwell angle is the angular displacement of the contact breaker shaft that determines how long the contact breaker remains closed. A correctly adjusted dwell angle is essential for two reasons: first, for correct timing of the sparks in the cylinders, and, with it, the highest possible engine efficiency; and second, for enabling the ignition coil to build up enough energy for the spark-over voltage.

The timing diagrams in Fig. 1b show how electrical pulses are obtained from the contact breaker. The top diagram shows the voltage typically developed across the contact breaker. This voltage is clipped and shaped to obtain digital compatible 5 V pulses that can be processed by the service module. The first negative pulse edge triggers a monostable multivibrator (MMV), which pulls its output low for a fixed period, T_{MMV}. The output of the MMV thus supplies a rectangular signal of which the 'low' time, T_L ($=T_{MMV}$), is constant in each period, while the 'high' time, T_H, is a function of engine speed: the trigger frequency rises with engine speed, while T_H becomes shorter. The average voltage, U_{av}, available at the output of the MMV is approximated with the equation

$$U_{av} \approx U_b T_H/(T_H+T_L)$$

Since the period of the contact breaker, T_0, is simply T_H+T_L, it follows that

$$T_0=1/f_0$$

where f_0 is the contact breaker frequency, which is a function of engine speed. From the above, it can be deduced that U_{av} is a function of engine speed:

$$U_{av}=U_b(T_0-T_{MMV})/T_0 = U_b[1-(T_{MMV}/T_0)] = U_b(1-f_0T_{MMV})$$

To understand how the dwell angle, Φ, is measured, it must first be defined as

$$\Phi=T(Tcl/T_0)(360/n)$$

where n is the number of cylinders.

A NAND gate is used to combine the shaped, digital signal (second drawing in Fig. 1b) with the MMV signal (third drawing). The result is the signal drawn in the last diagram in Fig. 1b. The combining is necessary to get rid of the noise at the start of each period of the input signal. The average value of the voltage

at the output of the NAND gate is written as

$$U_{av}=U_b[1-(T_L/T_o)]$$

Since, in a four-cyilinder, four-stroke, engine,

$$\Phi=90T(T_L/T_o)$$

it is evident that U_{av} is directly proportional to Φ, so that it can be used to measure the dwell angle.

Circuit description

Figure 2 shows that the circuit of the meter section of the service module is fairly simple, and essentially based on only one integrated circuit, the CMOS Type 4011. The 5 V regulator, IC_2, is fed from the 9 V battery in the display circuit described below. A zener diode, D_1, and a series resistor, R_1, reduce the amplitude of the contact breaker signal to a value suitable for applying to a CMOS NAND gate, N_1. Capacitor C_1 in the input network shunts any high-frequency components to ground.
Gate N_1 functions as a pulse shaper as already discussed with reference to Fig. 1b. Parts R_2 and C_2 form a differential network that supplies a very brief, active low, needle pulse with every negative pulse transition from N_1. The monostable multivibrator set up around N_2 and N_3 is triggered on the first of these needle pulses as shown in the timing diagrams in Fig. 3. In the non-triggered state, the MMV output (N_3 pin 10) as well as the input (N_2 pin 5) are logic high. Since the output is connected to pin 6 of N_2, pin 4 of this gate is logic low. This condition is stable with no voltage across C_3. Following a negative-going needle pulse at the input of the MMV, the output of N_2 toggles from low to high. The resulting charge current through C_3, shown in Fig. 3c, causes a quickly rising and a slowly, logarithmically decreasing, voltage drop across R_3 and P_1. Consequently, N_3 toggles: its output, and with it the second input (pin 6) of N_2, goes low, so that a stable situation is obtained for as long as the voltage across R_3 and P_1 does not exceed the toggle threshold of N_3.
When, at a voltage level of about $\frac{1}{2}U_b$, the input voltage of N_3 falls below the toggle threshold, the gate supplies a high level again. The monotime T_{MMV} is over, and both inputs of N_2 are logic high again. In other words, the stable stand-by state is restored until the next trigger pulse occurs.

VMOSFET T_1 blocks during the monotime. As soon as this ceases, however, the transistor conducts and effectively shunts P_1 and R_3 with a relatively low resistance, R_4. This causes C_3 to be discharged much faster, so that the monotime of the MMV remains constant even with relatively high trigger frequencies (= engine speeds). A VMOSFET Type BS170 is used here because its high input impedance ensures that N_3 is not overloaded. Moreover, the transistor has a very low drain-source saturation voltage, so that it does not affect the operation of in-

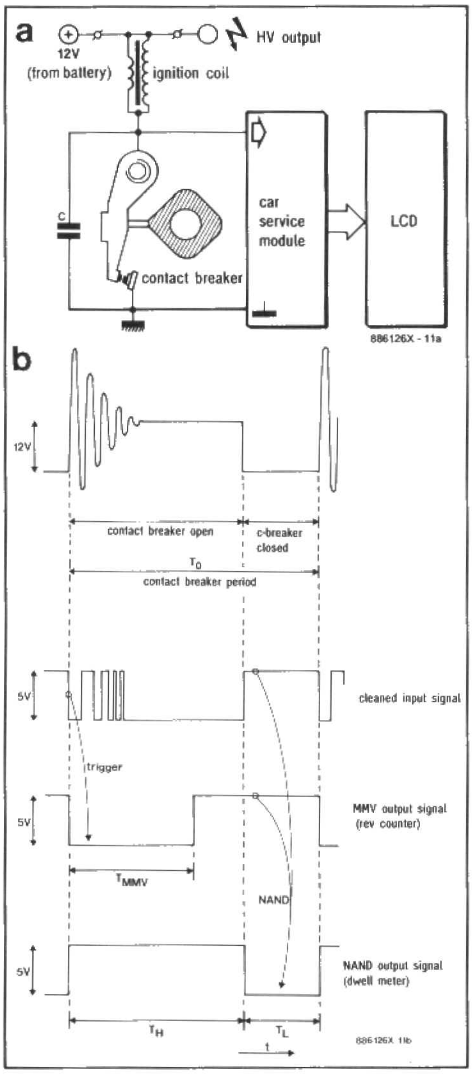


Fig. 1. Basic ignition system in a petrol-engine (1a) and timing diagrams of the car service module (1b).

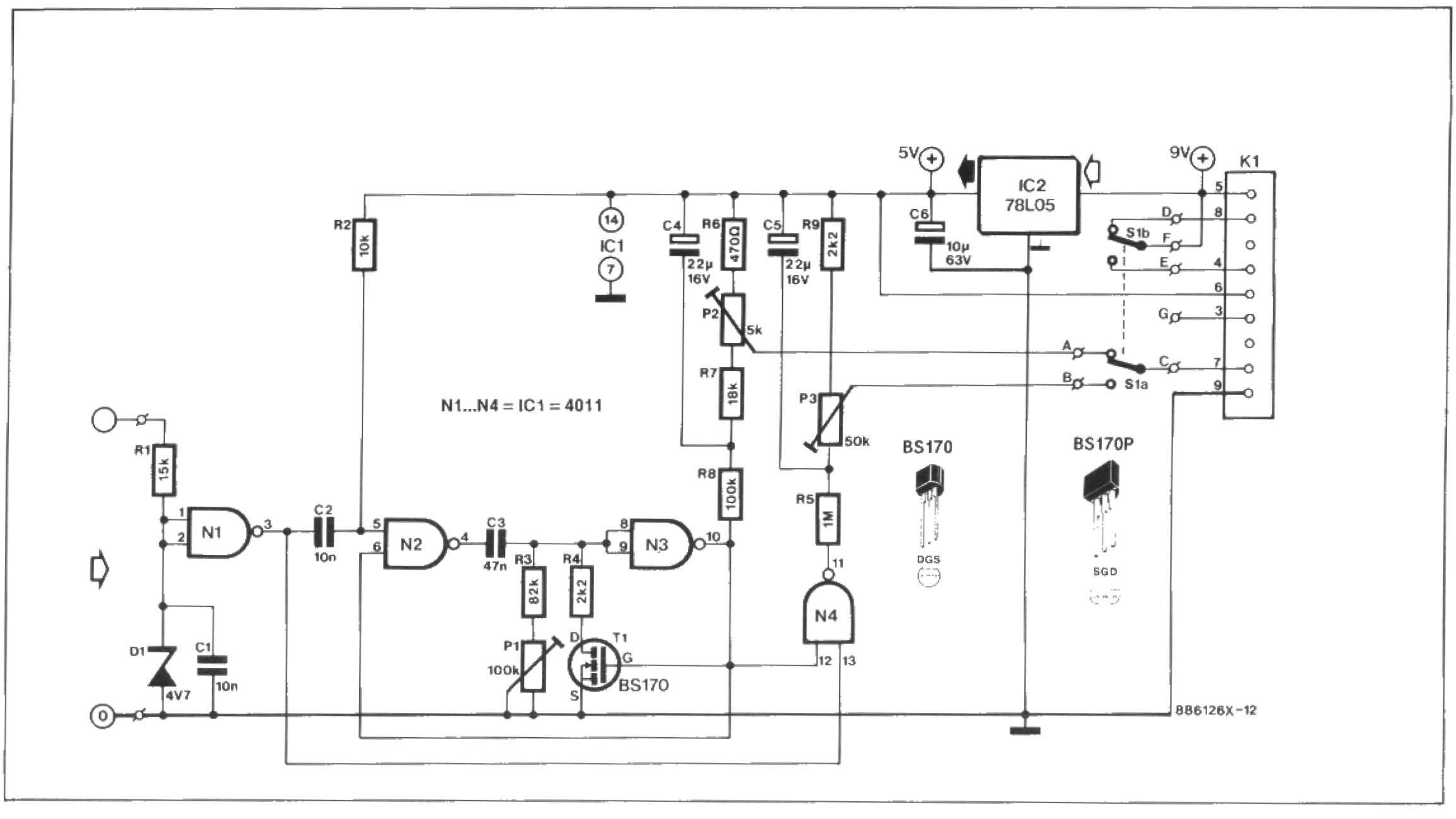


Fig. 2. Circuit diagram of the rev counter/dwell meter circuit in the car service module.

tegrator R_8-C_4. This network serves to convert the rectangular signal at the output of N_3 into a direct voltage that is directly proportional to the average voltage of the rectangular signal, and, therefore, to the engine speed. The capacitor, C_4, is connected to the positive supply line because U_{av} is obtained from the active-low output of the MMV, so that it is actually an *inverse* function of engine speed (refer back to Fig. 1b). With C_4 connected to the positive supply rail, this inversion is inverted again, since the voltage on the capacitor increases when U_{av} decreases.

Dwell angle measurement uses integrating network R_5-C_5 at the output of N_4. As shown in Fig. 1b, this NAND gate combines the cleaned input signal with the MMV signal, so that the voltage on C_5 is directly proportional to the dwell angle.
Finally, potential dividers R_6-P_2-R_7 (rev counter) and R_9-P_3 (dwell meter) provide the drive voltages for the LCD readout. The presets are used for calibrating the two functions of the module. Toggle switch S_{1a} selects between the revolution counter and the dwell angle meter functions, while S_{1b} selects the correct position of the decimal point on the display (DP2 for the rev counter, and DP1 for the dwell meter).

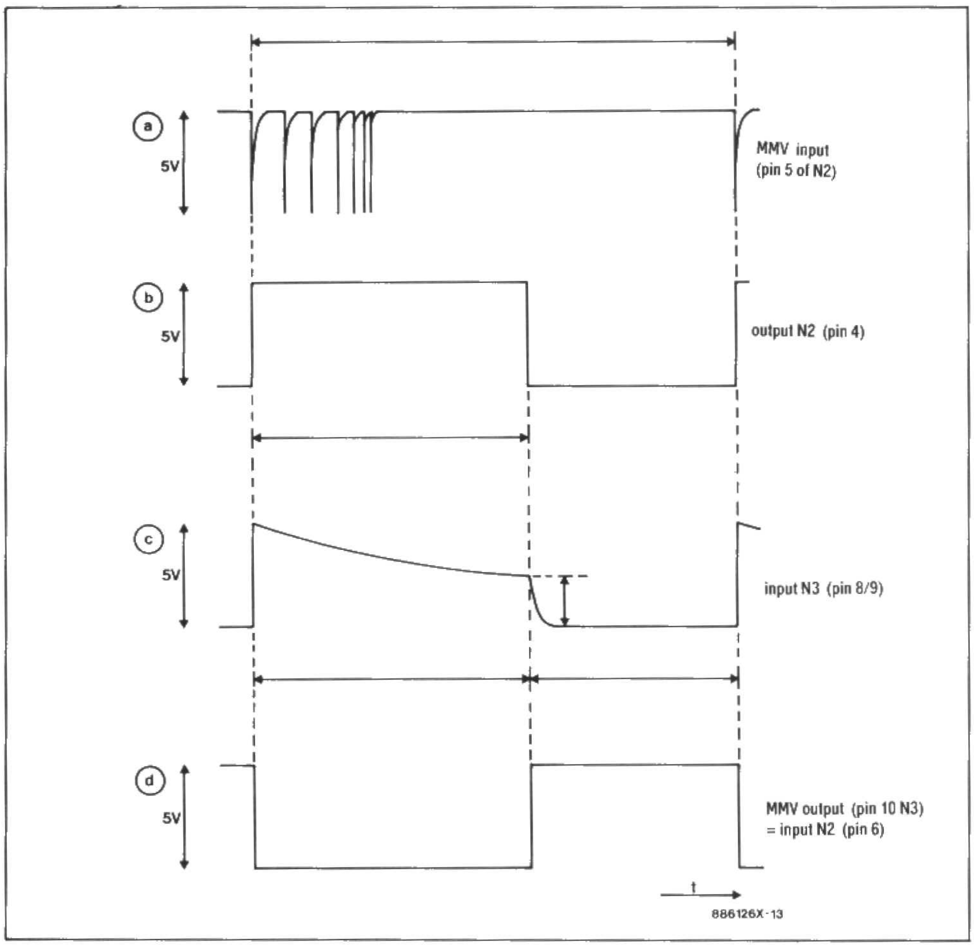


Fig. 3. Basic operation of the monostable set up around N_2 and N_3.

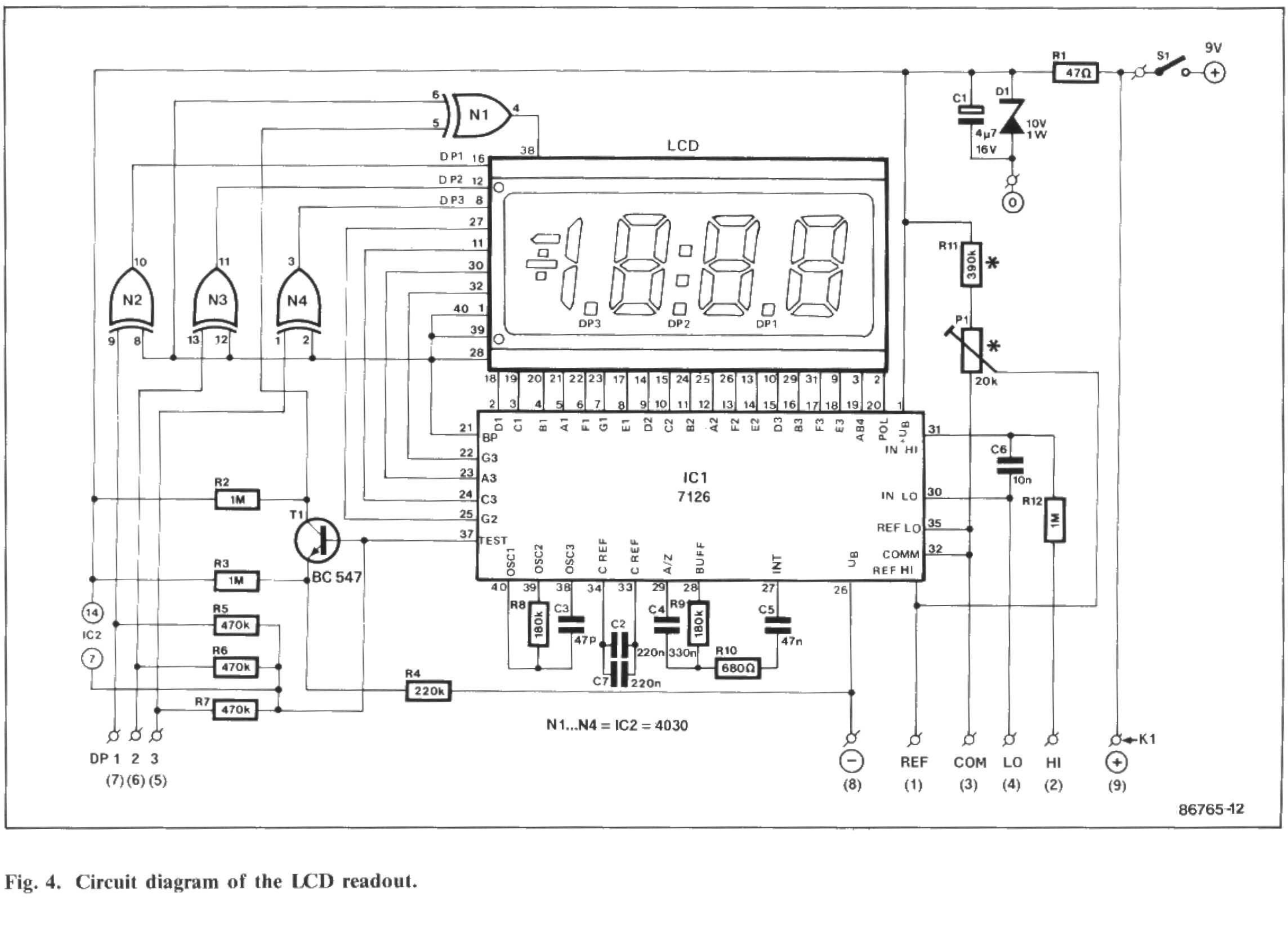
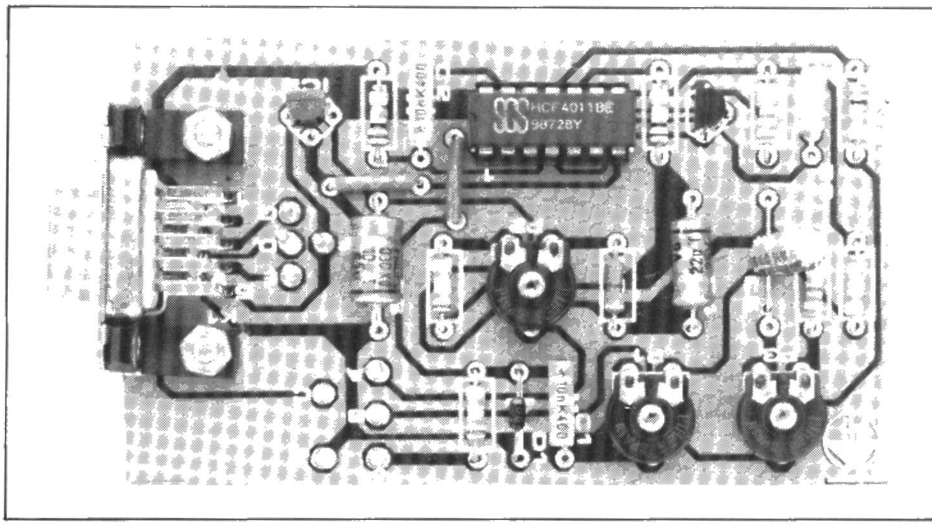


Fig. 4. Circuit diagram of the LCD readout.

A universal LC display module

The circuit diagram of Fig. 4 shows that the 3½-digit liquid-crystal display with the car service module is a standard application of the ICL7126 from Intersil (the ICL7126 is a CMOS version of the familiar ICL7106 which may also be used here). Transistor T_1 is added to actuate the LO BAT indicator on the display when the 9 V battery is exhausted (U_b<7.2 V). The auto-zero function of the ICL7126 obviates any null adjustments. The display unit is calibrated by interconnnecting its LO and COM inputs, applying a variable voltage between 0 and 200 mV to LO (−) and HI (+), and adjusting preset P_1 until the read-out is in accordance with the actual value of the applied voltage, which is measured simultaneously with a digital voltmeter.



Completed meter circuit on PCB 886126.

Construction and alignment

Building the two circuits that together form the car service module on the PCBs shown in Figs. 5 (meter section) and 6 (LC display) is straightforward. Angled 9-pin D-connectors for PCB mounting are used for interconnecting the circuits by means of a length of 9-way cable. The size of the PCBs is such that the units can be housed in identical, transparent, enclosures, from which the 9-pin connectors protrude. The input to the meter circuit is made by 2 wander sockets, a red and a black one, which accept plugs fitted on heavy-duty, heat-resistant test wires with insulated croc clips at the other end for connecting to the contact breaker on the car engine.

For the following description of the alignment of the service meter, it is assumed that the digital read-out has been calibrated as detailed above.

First, build the 50 Hz source shown in Fig. 7. The alternating voltage it supplies simulates the contact breaker pulses, and is applied to the input of the car service module. Since,in a four-cylinder, four-stroke, engine, ignition in a cylinder takes place every fourth revolution of the crankshaft, 50 contact breaker pulses per second simulate 50×60=3000 sparks per minute, or 750

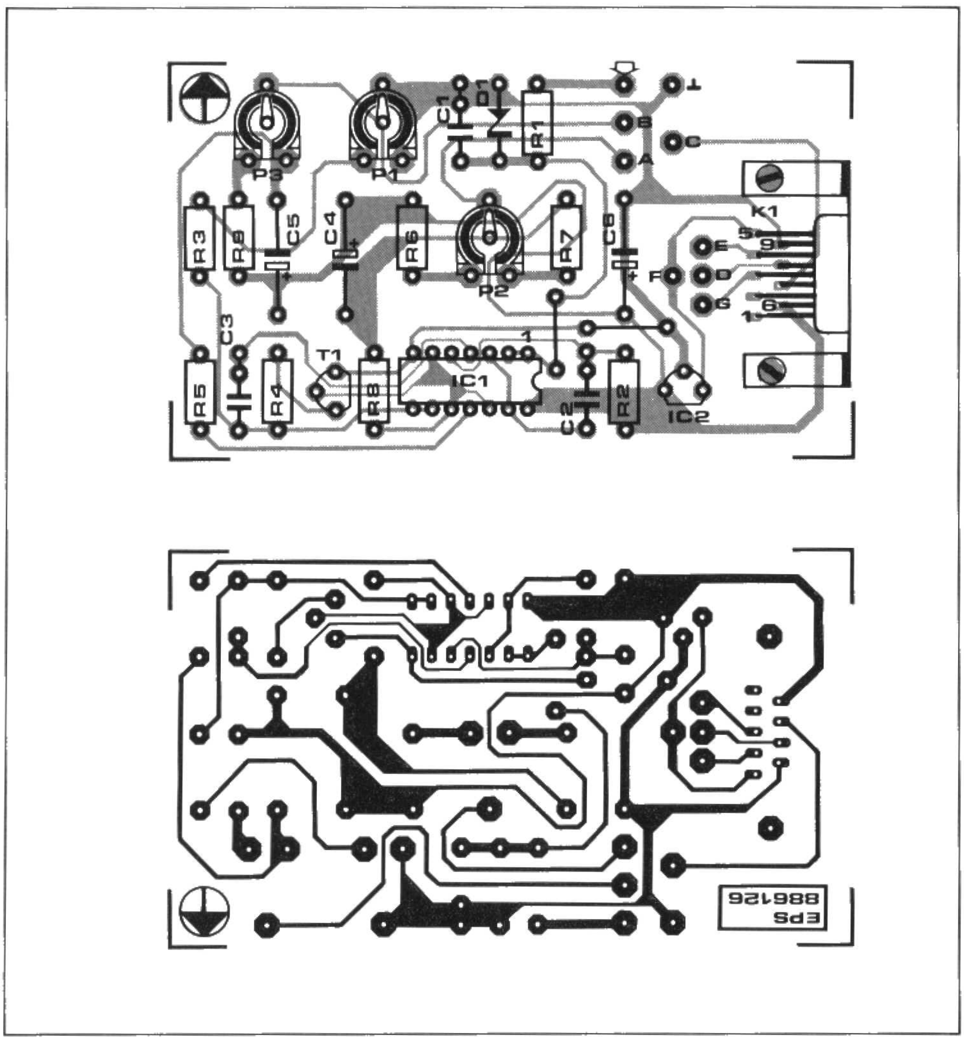


Fig. 5. Printed-circuit board for the meter circuit of the car service module.

Parts list

METER BOARD. CIRCUIT DIAGRAM: FIG. 2.

Resistors (±5%):
R_1 = 15K
R_2 = 10K
R_3 = 82k
R_4;R_9 = 2K2
R_5 = 1M0
R_6 = 470R
R_7 = 18K
R_8 = 100K
P_1 = 100K preset H
P_2 = 5K0 preset H
P_3 = 50K preset H

Capacitors:
C_1;C_2 = 10n
C_3 = 47n
C_4;C_5 = 22µ; 16 V; axial
C_6 = 10µ; 16 V; axial

Semiconductors:
D_1 = zener diode 4V7; 400 mW
T_1 = BS170
IC_1 = 4011
IC_2 = 78L05

Miscellaneous:
K_1 = 9-way male D connector for PCB mounting.
S_1 = miniature double-pole toggle switch (DPDT).
PCB Type 886126 (see Readers Services page).
Enclosure: e.g. Heddic Type 222 (Chartland Electronics Limited • Chartland House • Twinoaks • COBHAM KT11 2QW. Telephone: 037 284 2553).

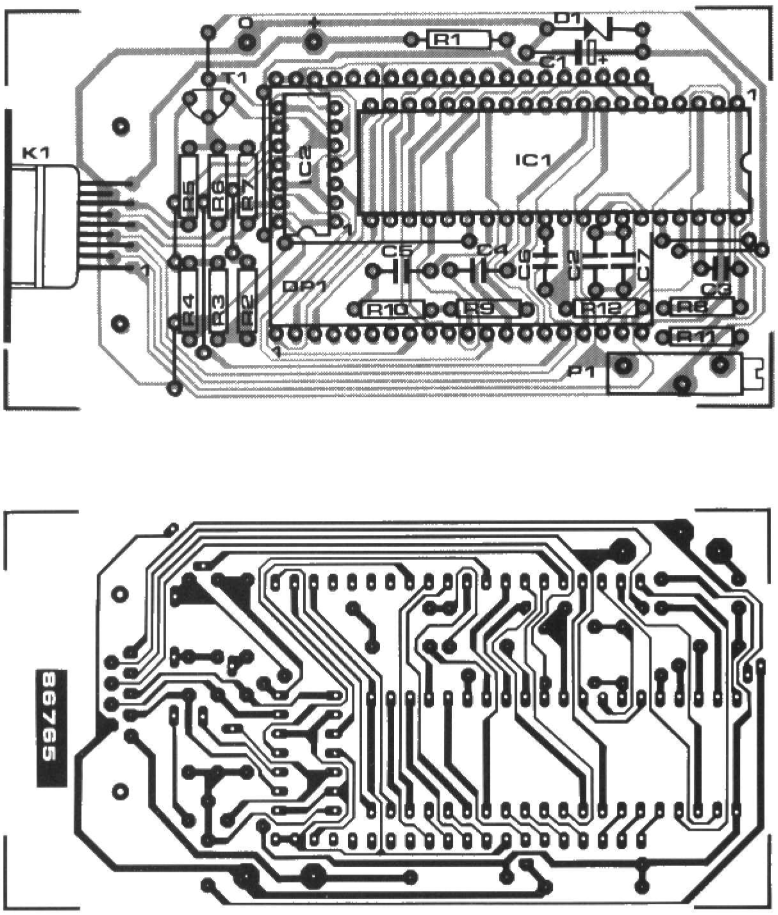


Fig. 6. Printed-circuit board for the LC display unit.

Parts list

LC DISPLAY UNIT. CIRCUIT DIAGRAM: FIG. 4.

Resistors (±5%):
R_1 = 47R
R_2;R_3;R_{12} = 1M0
R_4 = 220K
R_5;R_6;R_7 = 470K
R_8;R_9 = 180K
R_{10} = 680R
R_{11} = 390K

Capacitors:
C_1 = 4µ7; 16 V
C_2;C_7 = 220n
C_3 = 47p
C_4 = 330n
C_5 = 47n
C_6 = 10n

Semiconductors:
D_1 = zener diode 10 V; 1 W
IC_1 = ICL7126 (Intersil)
IC_2 = 4030
T_1 = BC547B

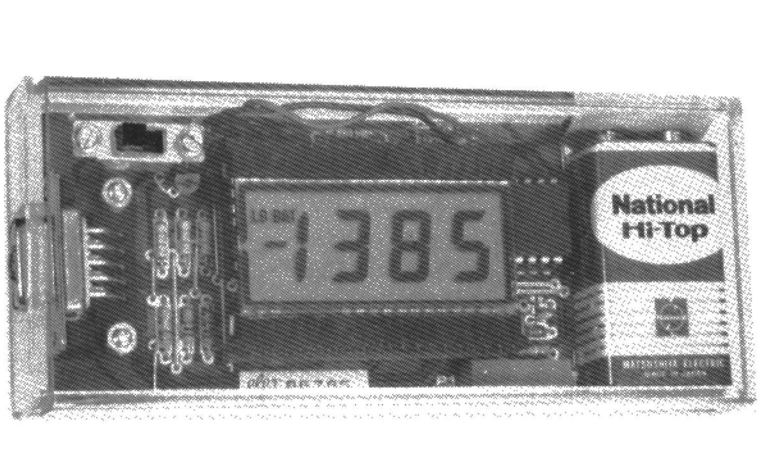
Miscellaneous:
LCD = general-purpose 3½-digit LC display with LO BAT indication.
K_1 = 9-way female D-connector for PCB mounting.
S_1 = miniature on/off slide switch.
PCB Type 86765 (see Readers Services page).
Enclosure: e.g. Heddic Type 222.

per minute per cylinder. Since one spark is genereated per two revolutions, the simulated engine speed is 1500 rpm. With a 60 Hz mains, this becomes 1800 rpm.

In most cases, the maximum engine speed will be about 6000 rpm, corresponding to a contact breaker frequency of 200 Hz. This means that the monotime of MMV in the circuit should be set to about 0.8(1/200) = 4 ms. Connect a high-impedance DVM across C_4, and adjust P_1 for a reading of 0.19 V. This sets the monotime with sufficient accuracy.

Make sure that the function switch, S_1, is set to rev counter, and adjust P_2 until the LCD readout indicates 1.5, which corresponds to 1500 rpm (60 Hz: 1800 rpm). Now switch on the dwell meter function and adjust P_3 for an LCD reading of 45.0°. In practical use of the instrument, it should be borne in mind that integrator R8-C4 is purposely dimensioned to give a stable readout, at the cost of a fairly slow meter response to engine speed variations. Also, since the input signal is combined with the MMV signal, dwell angle measurements can only be made at engine speeds lower than 3000 rpm.

The meter is also suitable for six-cylinder engines. Since these generally run at a lower speed than 4-cylinder types, no changes are, in principle, required to the previously detailed adjustment of P_1. The signal supplied by the test circuit of Fig. 7 then simply corresponds to 1000 rpm (60 Hz: 1200 rpm) and a dwell angle of 30°.



This compact, ICL7126-based LC display unit can be used in many applications.

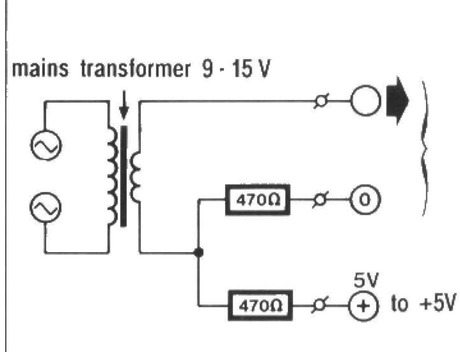


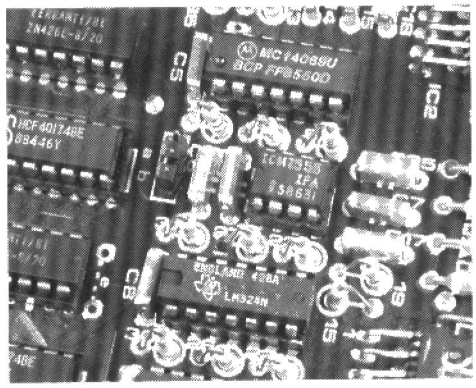
Fig. 7. A simple signal source for adjusting the car service module.

MORE APPLICATIONS FOR THE 555

There are probably few integrated circuits that have been with us for as long as timer Type 555. This article does not add to the seemingly endless list of AMV and MMV applications of this chip, but discusses some less familar designs derived from these. In addition, a brief introduction is given to the new CMOS and LinCMOS versions of the 555.

by T. Wigmore

One explanation of the popularity of the now 17-year-old timer type 555 may be that the chip is inexpensive, and contains a fairly unique combination of sub-circuits. Looking at the internal structure shown in Fig. 1, these are a bistable (a digital circuit), two comparators (analogue circuits) and some discrete parts, a resistive potential divider and a transistor. Added to the versatility of these interesting building blocks come the abilities of the chip to supply a relatively high output current, and to work from a wide range of supply voltages. Pin assignments of the 555 and the dual version of it, the 556, are given in Fig 2. Every electronic engineer or student is bound, at some time, to deal with the 555 in its standard configuration as a monostable or astable multivibrator. These applications are so numerous by now that it is often forgotten, or not even known, that the 555 can be used in a number of other, less well known, configurations. To understand how these work, however, it is useful to first look at the basic operation of the chip.



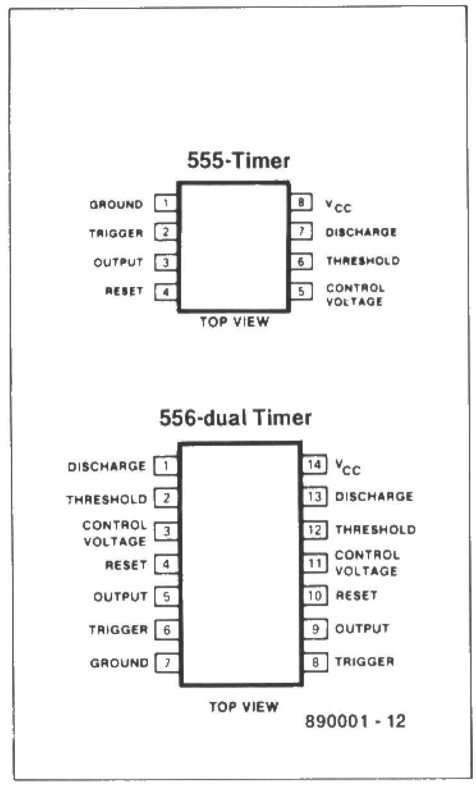


Fig. 2. Pinning of the 555 and the 556

Some fundamentals

Judging from the internal diagram of the 555 (Fig. 3), the relatively high number of components is typical of chip technology of the early 1970s. Fortunately, the internal diagram is still fairly simple to analyse. A trigger comparator (block A) and a threshold comparator, block B, are clearly recognized as difference amplifiers. The bistable, block C, is, perhaps, less conspicuous. In rest, transistors Q_{15} and Q_{17} are off, while Q_{16} and Q_{20} conduct. When the trigger voltage drops below one-third of the supply voltage, Q_{10}, Q_{11} and, therefore, Q_{15}, also start to conduct. Transistor Q_{15} removes the base drive of Q_{16} and so causes this to block. By virtue of R_{10} and diode Q_{18}, Q_{17} starts to conduct. As the trigger voltage rises again, Q_{15} is allowed to turn off again without causing instability of the new state — Q_{16} is

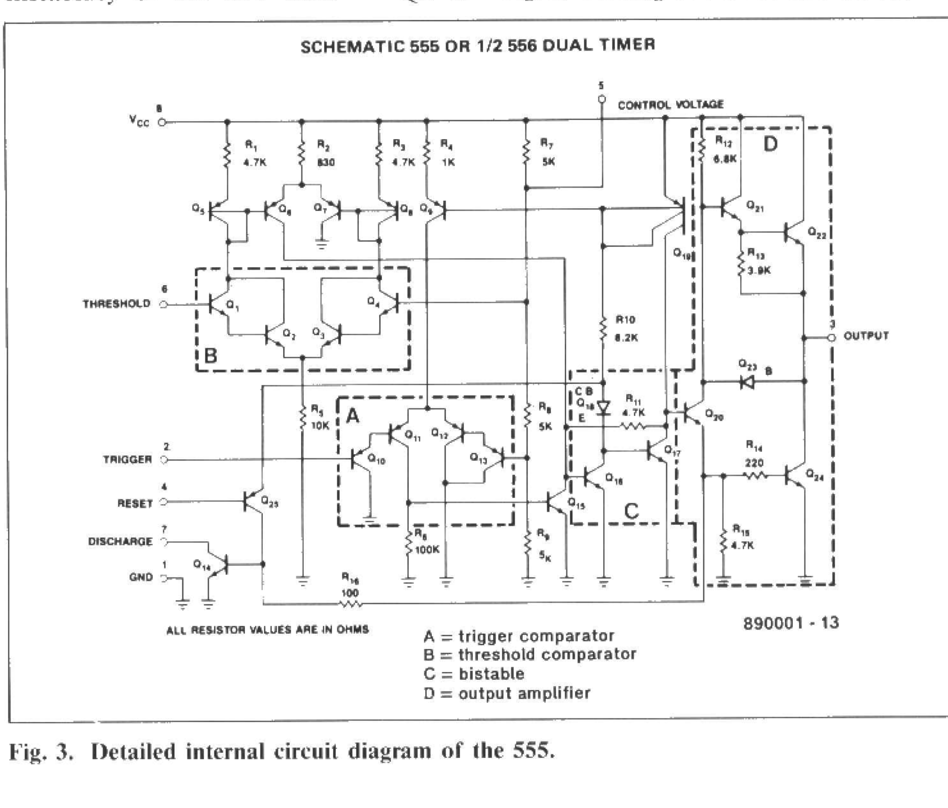


Fig. 3. Detailed internal circuit diagram of the 555.

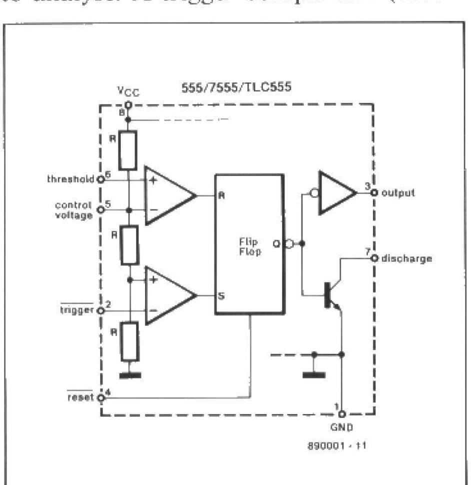


Fig. 1. Basic internal structure of the 555.

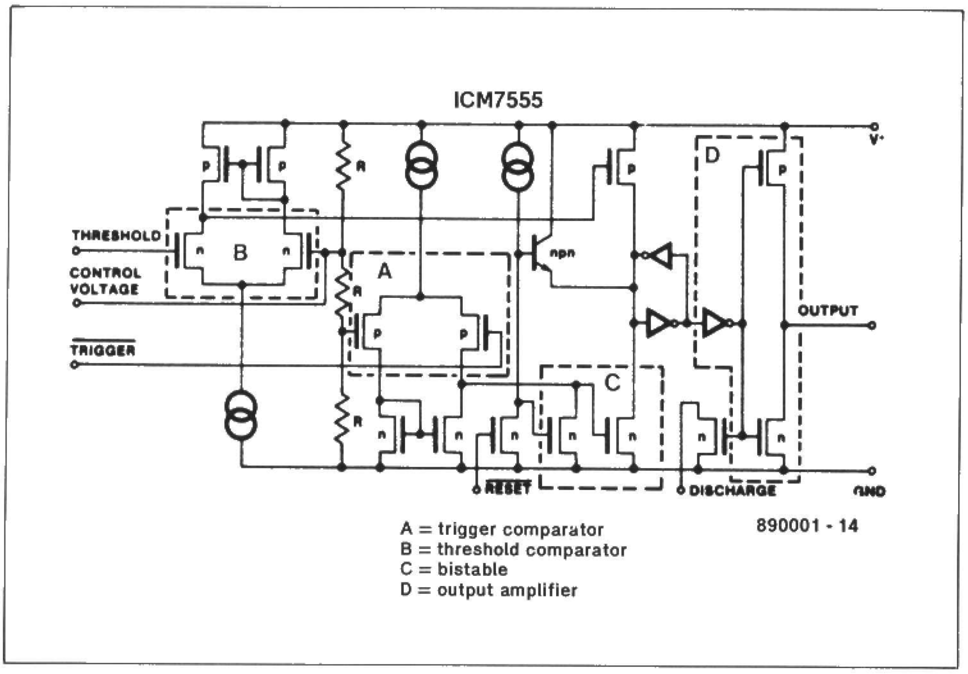


Fig. 4. Internal structure of a CMOS version of the 555, the ICM7555 from Intersil.

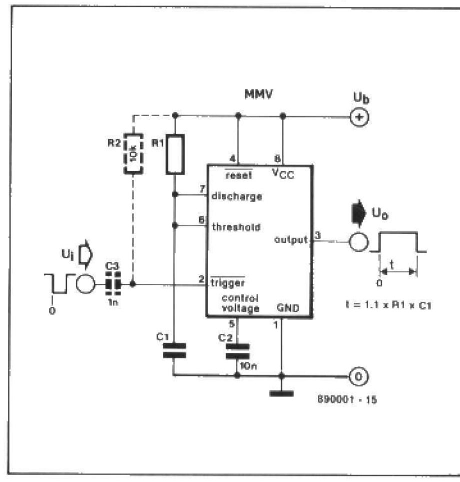


Fig. 6. Standard application of the 555 in MMV configuration.

a

b

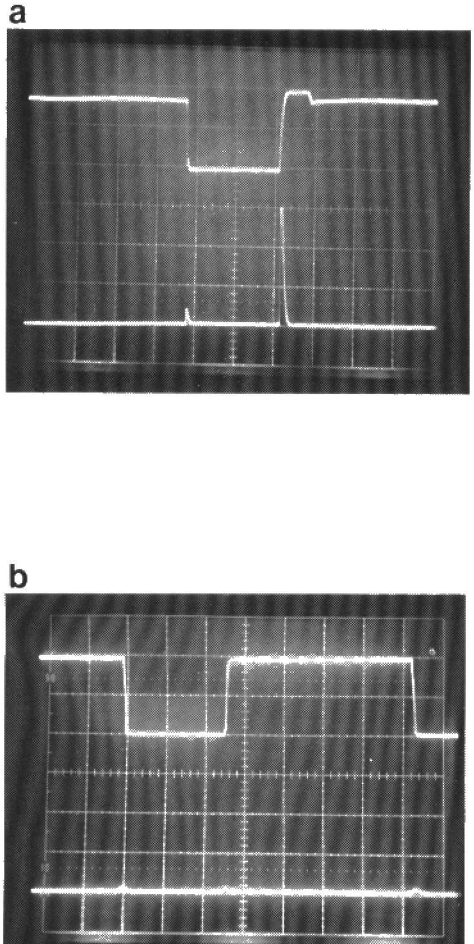


Fig. 5. A standard 555 briefly draws a high current when its output toggles (Fig. 5a; lower trace shows inverted supply voltage; decoupling of the supply voltage is a must!). The new CMOS 555 does not produce this annoying effect (Fig. 5b)

then inhibited from conducting via R_{11}. The normal procedure is that the threshold voltage exceeds two-thirds of the supply voltage. This results in Q_1 and Q_2 starting to conduct. The increase in their collector currents is amplified by Q_5 and Q_6, so that Q_{16} starts to conduct again. This transistor, in turn, causes Q_{17} to block, but only if Q_{15} is actually off. If this is not so — in other words, if the threshold input and the trigger input are both actuated — the bistable remains reset. Because the collector current of Q_6 is limited by R_2, Q_{15} pulls the base of Q_{16} harder to ground than Q_6 can pull it to the positive supply rail.
An all-overriding method to reset the bistable is to drive its reset input low. This results in Q_{25} conducting, so that the base drive of Q_{17} is removed. Since diode Q_{18} creates additional voltage drop during resetting, the base voltage of Q_{17} is sufficiently low to actually turn this transistor off. When the bistable is in the reset state, output transistors Q_{20} and Q_{24} and, via R_{16}, discharge transistor Q_{14}, conduct.

The 555 briefly draws a fairly high current when its output changes from low to high. This is so because Q_{24} is briefly driven into saturation, and takes a while to actually turn off. As soon as Q_{21} and Q_{22} conduct, a short, non-current limited, short-circuit of the supply arises. It is for this reason that the 555 requires particular attention to be paid to decoupling of the supply voltage (see Fig. 5a). Output switching from high to low causes fewer problems because Q_{21} and Q_{22} are not driven into saturation; hence, the switch-off time is short relative to that of Q_{24}. CMOS versions of the 555 generally do suffer from this annoying effect.

Applications

In 9 out of 10 applications of the 555, the chip is used as a monostable or astable multivibrator (AMV or MMV respectively). In MMV configuration, the pulse time is determined by the time needed to charge the timing capacitor from 0 V to $\frac{2}{3}U_b$, the *threshold voltage*. In general, the charge voltage, U_c, on a capacitor, C, charging through a resistor R, from a supply voltage, U_b, is equal to $\frac{2}{3}U_b$ when

$$U_c(t)=U_b(1-e^{-\tau/RC})$$

from which,

$$\tau=(-\log_e{}^1\!/\!_3)RC=1.1RC$$

The charge voltage also determines the monotime, provided the trigger pulse is shorter than the monotime. A longer trigger pulse also results in a longer output pulse, but this may be prevented by driving the trigger input with an AC-coupled signal only (add R_2/C_3, with $(R_2C_3)<(R_1C_1)$).

The MMV circuit is turned into an AMV simply by making it self-triggering. Capacitor C_1, via R_1 and R_2, is charged to $\frac{2}{3}U_b$ in time interval t_1:

$$t_1=(-\log_e{}^1\!/\!_3)(R_1+R_2)C-(-\log_e{}^2\!/\!_3)(R_1+R_2)C$$
$$=0.694(R_1+R_2)$$

and is then discharged again, this time only via R_2. The discharge time, t_2, equals

$$t_2=0.694R_2C$$

This means that the voltage on the capacitor toggles between $\frac{1}{3}U_b$ and $\frac{2}{3}U_b$. The total period, T, is calculated as

$$T=t_1+t_2=0.694(R_1+2R_2)C$$

and the frequency, f_0, as

$$f_0=1/T=1.44/(R_1+2R_2)C$$

It should be remembered, however, that C_1 has to be charged from 0 V when power is first applied, or when the reset input is made high. The first part of the first output period, therefore, has a period of $1.1R_1C_1$.

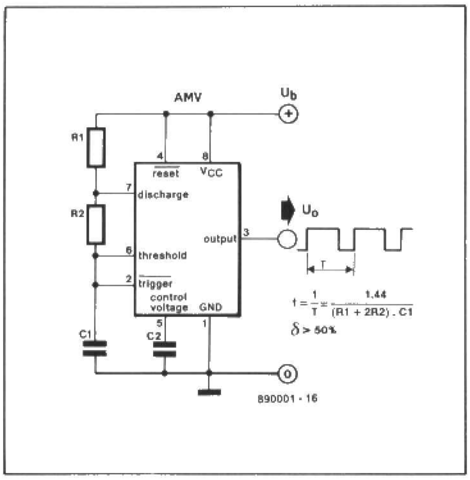


Fig. 7. Standard application of the 555 in AMV configuration.

One of the nice features of the 555 as an MMV or AMV is that the pulse time is, in principle, independent of the supply voltage, U_b. When this drops, the trigger and threshold voltages, as well as the charge- and discharge currents, drop accordingly, resulting in no change overall. A disadvantage of the AMV circuit shown in Fig. 7 is its inability to supply an output signal of duty factor greater than 0.5: this is because the charge resistance, R_1+R_2, is always greater than the discharge resistance, R_2 by itself. The basic circuit in Fig. 8 shows how this can be resolved with the aid of a diode, D_1. During charging, it bypasses R_2, so that the charge current can become smaller than the discharge current. Another diode, D_2, is optional if R_1 alone is to determine the charge current. It should be noted that the above use of diodes sacrifices, at least partly, the 555's independence of the supply voltage level — when the supply voltage is changed, the fixed drop across the diode results in a non-proportional change of the charge and discharge current of C_1.

The control voltage input, pin 5, of the bipolar 555, is normally decoupled to ground with a 10 nF capacitor for noise protection. According to the manufacturers, this capacitor is no longer required with the new CMOS versions of the 555.

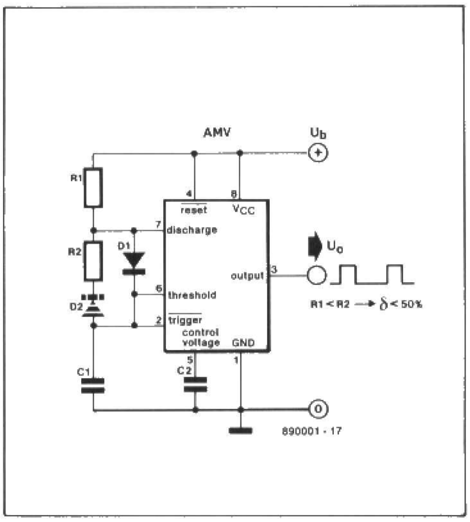


Fig. 8. Non-standard AMV configuration that allows duty factors lower than 0.5 to be achieved.

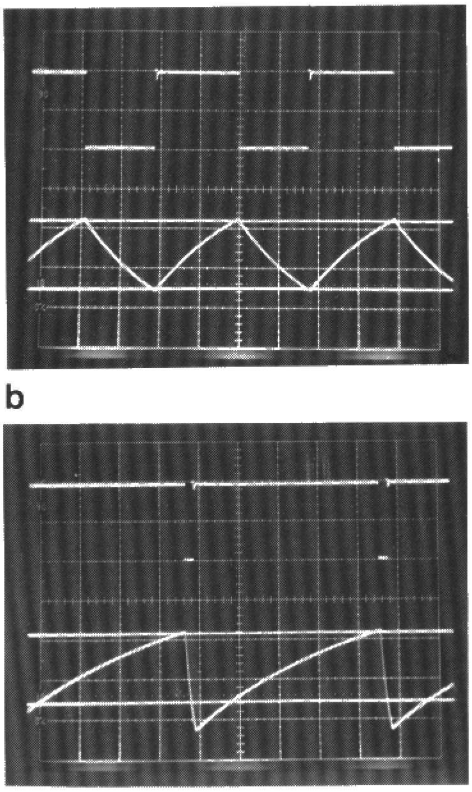


Fig. 9. Frequency deviation of a 555 in AMV configuration is a function of a number of parameters, including the duty factor. The effect shown by these oscillograms is mainly on account of the recovery time of the trigger comparator and discharge transistor. Upper trace: output signal; lower trace: voltage on timing capacitor. The horizontal traces show the trigger and comparator threshold levels.

Timing errors

It is not so simple to express the inaccuracy of a timing interval produced by a 555 as a single error-percentage. A large number of factors should be taken into account here, but many can be forestalled by correct dimensioning and/or selection of the most appropriate type of 555 for a particular application. Tolerance on the internally generated reference voltages, in combination with input-offset voltages of the trigger- and threshold comparators, introduces timing errors of the order of 2%.

Internal reaction and recovery times also form a factor to be taken into account. The oscilloscope photographs in Fig. 9 illustrate the behaviour of a 555-based AMV at a relatively high output frequency. Figure 9a shows the AMV set to a duty factor of about 0.6. The frequency, 29 kHz, already deviates considerably from the calculated 25 kHz. Fig. 9b shows the output signal of the same circuit, this time dimensioned for a much greater duty factor. Since the total resistance R_1+2R_2 is equal in both cases, it might be expected that the output frequency remains unchanged. It is seen, however, that C_1 is actually discharged to below the trigger level (which, like the threshold level, is marked by a horizontal trace). This effect is caused partly by the relatively quickly falling voltage on C_1, and partly by the slowness of the trigger comparator in combination with the recovery time of the discharge transistor. Because of the excess discharge of C_1, the output frequency of the 555 will be significantly lower than calculated: 20 kHz in this case.

The essence of all this is that the accuracy of relatively high output frequencies depends largely on the duty factor.

When the 555 is configured as an MMV, due account should be taken of the saturation voltage of the internal discharge transistor. The level of this saturation voltage is inversely related to the value of the charge resistor, and, at relatively short monotimes, causes the output pulse to be shorter than calculated.

At very low output frequencies, factors such as the leakage current of the timing capacitor, that of the discharge transistor, and the input current of the threshold comparator, become increasingly significant.

In general, the lower the frequency, the higher the values of the charge and discharge resistors. As the charge current decreases, the importance of various leakage currents increases. Also remember that the use of an electrolytic capacitor with high leakage and tolerance in position C_1 will cause a much higher timing error.

Using the control input

The control voltage input, pin 5, affords a number of interesting, yet little used,

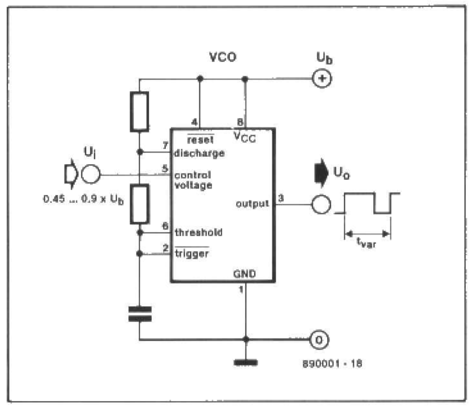


Fig. 10. By using the control voltage input, a 555-based AMV can be turned into a VCO.

applications, whose background is discussed below.
The internal diagram shows that pin 5 is connected to the internal voltage divider. When not externally loaded, this carries a voltage of $\frac{2}{3}U_b$. According to the manufacturers, this voltage may be varied between 45% and 90% of the supply voltage. When the control voltage is made too high, however, the threshold comparator will cease to work correctly, while a too low voltage at the control input upsets the bias point of the trigger comparator (refer to the internal diagram in Fig. 3).
The most evident application of the control voltage input is, of course, the 555 as a voltage-controlled oscillator (VCO), as shown in Fig. 10. The 555 itself is configured as an AMV whose output frequency can be varied over about ±50%. In practice, especially when the supply voltage is relatively high, a value considerably lower than $0.45U_b$, but with a minimum of about 1.5 V, is permissible for the control voltage. The frequency so achieved becomes up to $2f_0$.

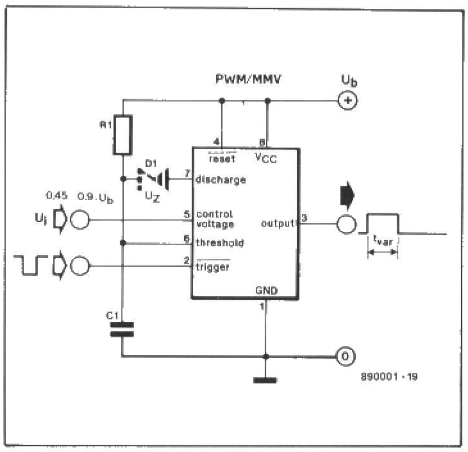


Fig. 11. Voltage-controlled monostable multivibrator.

The basic circuit of Fig. 11 shows that the control voltage input may also be used for making an MMV with adjustable monotime. When, however, the standard monostable configuration is chosen, the output pulse can never become too short. Assuming an input voltage, U_i, at pin 5 of $0.45U_b$, the voltage on C_1 will be kept at virtually 0 V by the internal discharge transistor. When a relatively large control range of the output pulse is desired, the lowest voltage on C_1 may be raised with the aid of a zener diode, or a number of series-connected, forward-biased, diodes, in the collector line of the discharge transistor. To obtain a well-defined minimum voltage on C_1, the quiescent current through R_1, I_{R1}, must be just high enough to achieve the correct zener voltage, U_z. This current is calculated from

$$I_{R1} = (U_b - U_z)/R_1$$

In practice, a few mA will suffice to achieve the zener effect.

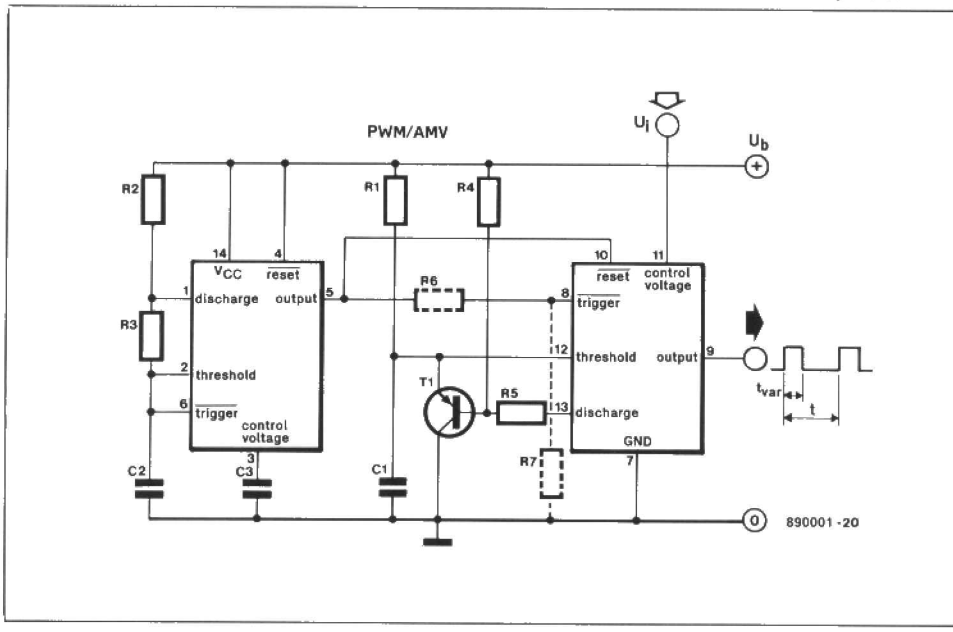


Fig. 12. Two 555's, or, in this case, a single 556, make an excellent fixed-frequency pulse-width modulator for low-loss power control systems.

The circuit of Fig. 11 does not provide a linear relationship between control input voltage and output pulse-width. Such linearity can be achieved, however, by replacing R_1 with a current source. A practical example and a detailed explanation of this interesting configuration is given in Ref. 1.
It is fairly simple to change the basic voltage-controlled monostable into a pulse-width modulated oscillator — see Fig. 12. All that is required is another AMV-based oscillator, set up around the other 555 contained in the 556 chip. The resulting circuit is an excellent, low-loss, pulse-width modulator for use with a power-transistor driver stage.
There are a few more interesting details in the circuit shown in Fig. 12. The first has to do with C_1, which is not discharged to 0 V, but to a level set with p.d. R_4-R_5, plus the base-emitter drop of T_1. Similar to the previously discussed 'zener-trick', this arrangement considerably magnifies the span of the output pulse-width.
The second interesting point of the circuit entails the simultaneous resetting and triggering of MMV_2 to ensure an accurately defined voltage on C_1 at the start of the each period. In the absence of the trigger signal, a curious phenomenon would take place when the duty factor is, theoretically, as close as possible to 1. During the first period, the threshold voltage is not reached, so that C_1 is not discharged. Immediately after the start of the second period, however, the threshold level is reached, so that the output goes low. The result of this sequence would be the halving of the output signal frequency, and a reducton of the duty factor from almost 1 to about 0.5.
As already said, this effect is prevented by resetting the MMV at the start of each period. Referring back to the internal diagram, the bistable is actually set and reset at the same time. Reliable triggering is, however, still ensured by virtue of the internal reset circuit switching off faster than the trigger circuit (Q_{15} has been driven into saturation, and has a longer recovery time). Incidentally, the recovery time of the trigger circuit can be shortened by using a potential divider that provides a trigger level just lower than $\frac{1}{3}U_b$.
In the concept discussed here, the duty factor can never become 1, because the output is invariably low for the duration of the reset signal of the MMV. This is why R_3 is generally made small relative to R_2.
The control voltage input of a standard 555 forms a fairly low resistance (5 kΩ//10 kΩ = 3.3 kΩ typ.). CMOS versions of the 555 have a much higher input resistance thanks to an internal voltage divider composed of three 100 kΩ resistors. In general, tolerance on these input resistance values is relatively high, so that a voltage source driving the control input should be designed to have a low output impedance.

Long-interval timers

As already hinted at in the section on timing errors, configuring the 555 as a long-interval timer may pose problems because of the inevitable role of leakage currents in the timing components, i.e., the high-value resistor(s) and the capacitor. A further aggravating effect is that the leakage current of an electrolytic capacitor is age- and temperature-dependent. In practice, the maximum interval that can be achieved with a 555 in standard configuration is 10 to 30 minutes long, taking a fairly high tolerance for granted.

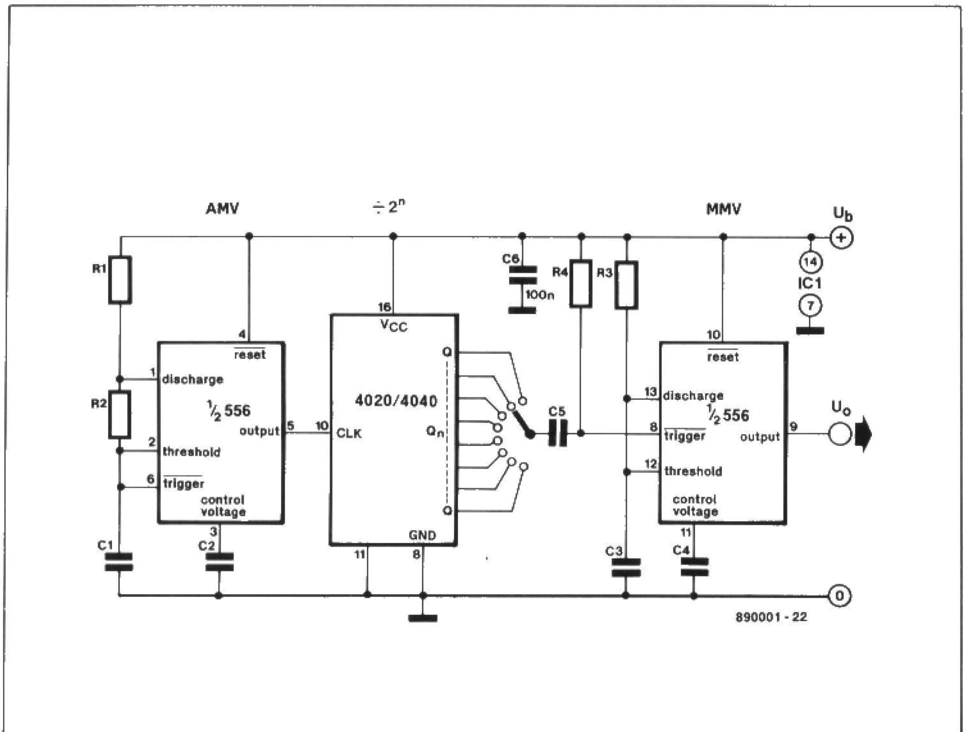


Fig. 13. Long-interval timers are best realized with the aid of a ripple-cascade divider.

One solution to obtain better-defined and longer intervals would be the cascading of 555s, so that each is triggered by the previous one. This is not a very neat solution to the problem, however, since all timing errors of individual timers in the cascade simply add up (accumulation effect). Moreover, the duration of the interval rises only linearly with the number of 555 stages. The increase can be made exponential by following one 555 in AMV mode with a divider as shown in Fig. 13. Depending on the application, the *n*-th output of the divider can trigger a further 555, this time in MMV mode. In this set-up, the 555 in AMV mode is conveniently dimensioned for optimum accuracy (average values for R_2 and R_3, and a low-leakage capacitor for C_1), while cascaded dividers afford timer intervals of hours, days or even weeks.

CMOS versions: 7555 and TLC555

Intersil was the first to introduce the 7555, a CMOS version of the 555. A little later, Texas Instruments, in line with its consistent and successful policy of producing LinCMOS (linear CMOS) versions of 'bipolar bestsellers', came up with the TLC555. As with a number of well-established opamps and comparators, the TLC555 and TLC556 from TI were an instant success.
In general, current consumption of the CMOS versions has been drastically reduced with respect to the bipolar 555 — from 10 mA to 100 μA, while the minimum supply voltage has been lowered to 2 V. Obviously, these features are of great importance for the design of battery-powered circuits. The CMOS versions do not suffer the large peak current at output switch-over, while the input bias current of the threshold comparator, and the leakage current of the discharge transistor, are also significantly reduced. These features of the new devices are advantageous because they allow a higher charge resistance for the capacitor, bringing longer timing intervals within reach.
Thanks to the virtual absence of saturation effects commonly associated with bipolar transistors, speed of the new CMOS 555's has also increased. In a laboratory test, a standard 555 gave up at about 180 kHz, whereas a 7555 scored 1.1 MHz, and a TLC555 even 2.4 MHz (test conditions: AMV configuration with $R_1=R_2=220\ \Omega$ and $C_1=100$ pF).
As far as output current is concerned, however, the bipolar 555, with its sink and source capabilility of 200 mA, is still superior to the CMOS versions. The 7555 supplies a maximum of 5 to 50 mA, depending on the supply voltage (10 mA at 10 V). The TLC555 has a symmetrical output with a source and sink capability of 10 mA and 100 mA respectively. *Ergo*, where the replacement of a standard 555 with a CMOS type is considered, the current requirement of the load should be taken into account (a standard 555s is often used to power a relay direct).

Reference:

1. Long-range infra-red transmitter-receiver. *Elektor Electronics* November 1987; p. 40 and 41.

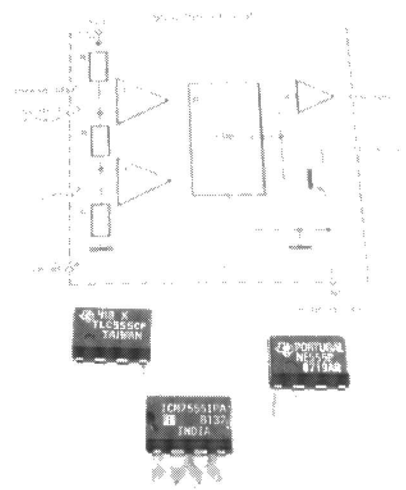


Table 1

		555			7555			TLC555			
		min.	typ.	max.	min.	typ.	max.	min.	typ.	max.	unit
V_{cc}/V_{dd}		4.5		18	2		18	2		18	V
Supply current	2V		-	-	-		-			0.25	mA
	5V		3	5	0.08		0.4		0.17	0.35	mA
	10V		10	12	0.12		0.6		0.36	0.60	mA
Output Current	I_{sink}		200		8		80		100		mA
	I_{source}		200		1		20		10		mA
Threshold current			100	250			10		0.01		nA
Discharge state-off current			20	100			10		0.1		nA
MMV timing error			1	3		2			1	3	%
Temp. drift				500		250			-		ppm/°
V_{cc} drift Output				0.5		0.3	1		0.1	0.5	%/V
rise-time			100	300		75			20		ns
fall-time			100	300		75			75		ns
f_{max}				0.5		1			2		MHz

data valid at $T_a = 25$ °C.

DESIGN IDEAS

The contents of this column are based solely on information supplied by the author and do not imply practical experience by *Elektor Electronics.*

SPEED CONTROL FOR ASYNCHRONOUS MOTORS

The possible design is discussed of a remarkably simple frequency converter that ensures good torque over a wide speed range for mains-powered motors of up to 100 W. Applications of the speed controller include ventilators and pumps.

by K. Walters

Induction motors with a squirrel-cage rotor are popular because of their simplicity and low cost. Compared with other types of AC-powered motors, speed control for the squirrel-cage induction motor, is, however, a relatively complex matter. Simply reducing the motor voltage generally gives poor speed control, and results in inacceptable loss of torque at low speeds. Frequency control of the motor supply voltage is a better method, since it exploits the fact that speed of the squirrel-cage motor is directly related to frequency. The basic requirements of a frequency controller circuit may be defined after briefly discussing the operating principles of the squirrel-cage motor.

A revolving transformer

The rotor in a squirrel-cage induction motor is of remarkable rubustness and simplicity. The windings of the rotor are formed by conduction bars, connected at either end by a short-circuiting ring (see Fig. 1a). In a practical construction, this cage-like structure is encapsulated in a tin-plate cover to keep the distribution of the magnetic field under control. The short-circuiting rings are, however, nearly always visible when the motor is dismantled.

The cage-type rotor is placed in a stator with two sets of poles. Figure 1b shows this structure for a single-phase motor. One set of poles holds the main winding, the second the auxiliary winding. By applying the mains voltage direct to the main winding, and, phase-shifted by 90°, to the auxiliary winding, the two

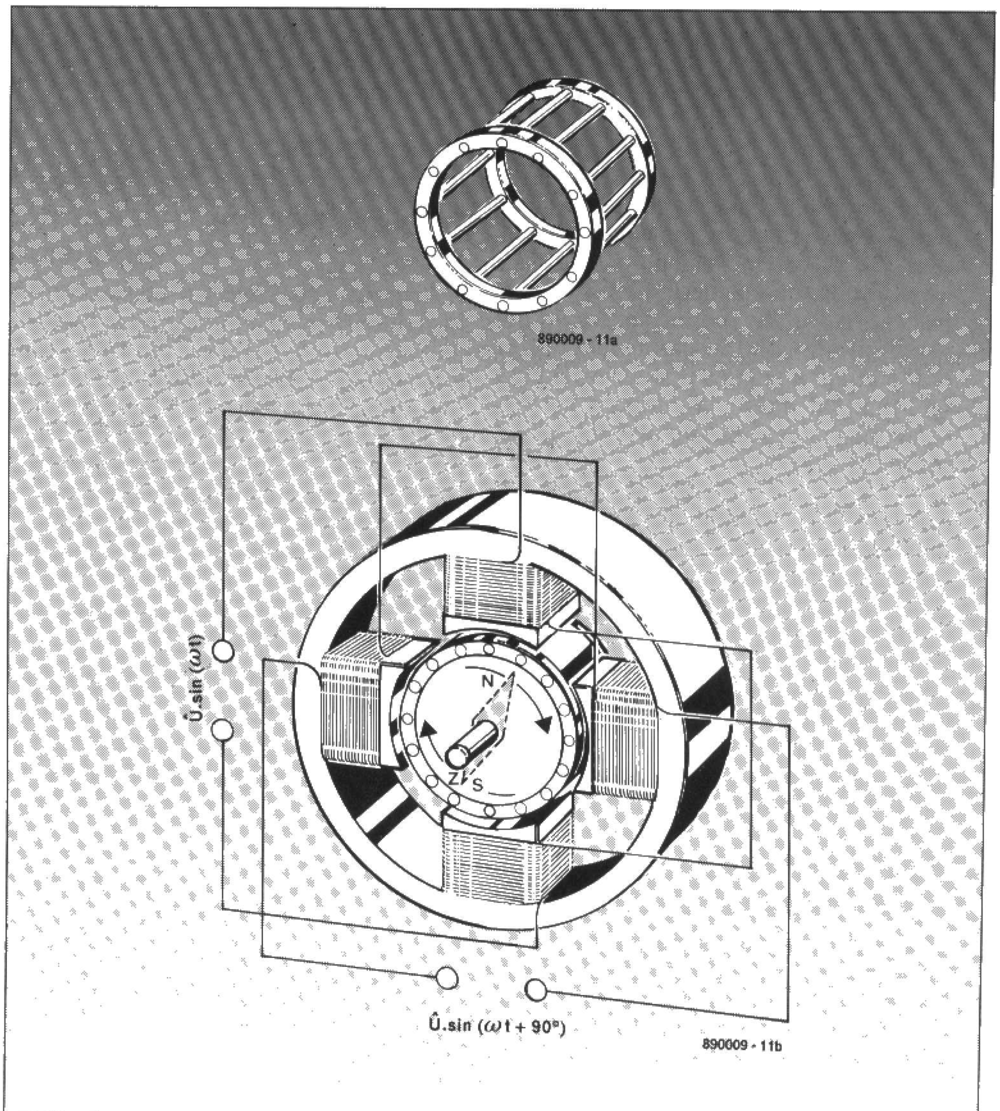


Fig. 1. Basic structure of the squirrel-cage induction motor: rotor (1a) and cross-sectional view of the stator and rotor (1b).

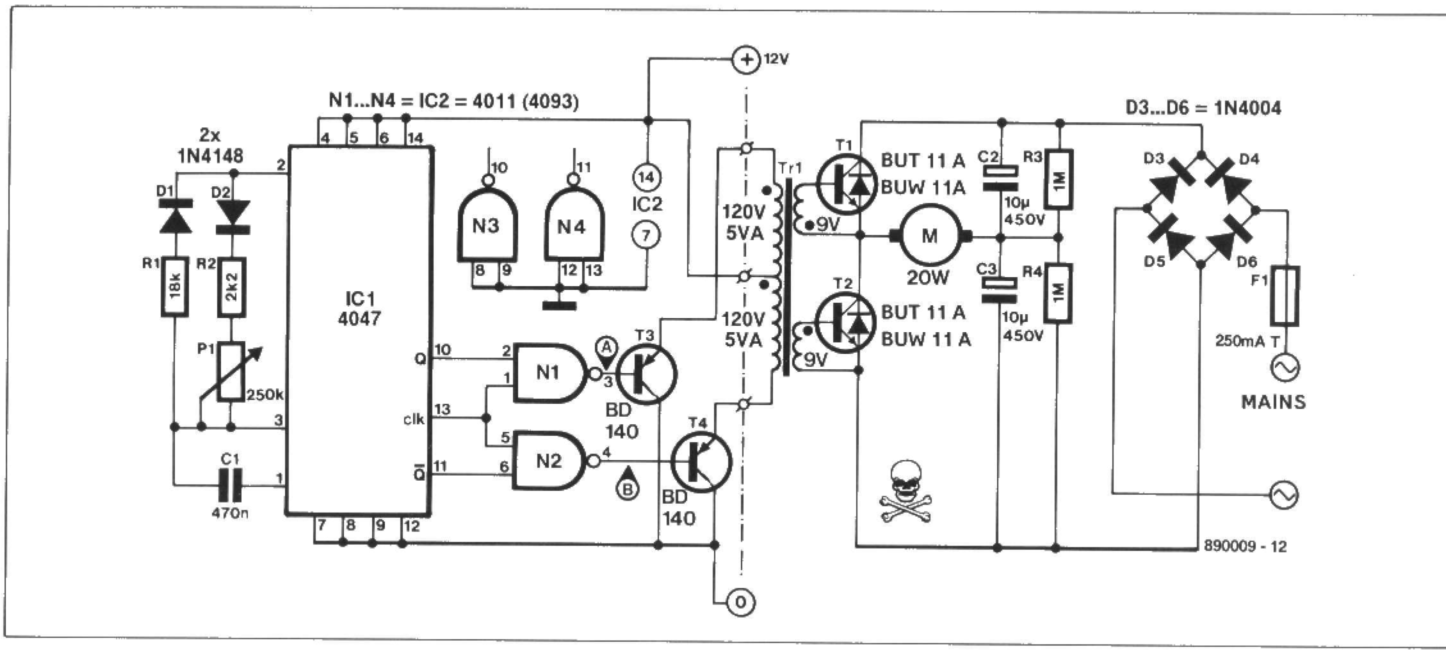


Fig. 2. Circuit diagram of the speed controller for asynchronous motors.

sets of poles create a rotating magnetic field. As in a transformer, a voltage is induced in the conduction bars of the cage. The ends of the bars are, however, commoned, giving rise to a high current that, in turn, generates a magnetic field. This causes the cage to be drawn along with the rotating magnetic field provided by the stator. Because the motor starts to run in the direction of travel of the stator field, the speed at which the rotor 'sees' the magnetic field revolve, is reduced. This effectivly reduces the current in the conduction bars, and, with it, the magnetic field strength of the cage. In practice, the actual movement of the cage lags that of the magnetic field by a few percent, so that motor speed is not synchronous with the stator field or, for that matter, the mains frequency. Hence, the type of motor discussed here is sometimes referred to as an *asynchronous* type.

The auxiliary winding is, in principle, only required to overcome the inertia of the cage when the motor is started. In some motors, the supply to this winding is interrupted by a centrifugal switch. Unfortunately, this type of motor can not be used with a speed control circuit of the type discussed here, because the centrifugal switch remains closed at relatively low speeds and would cause the auxiliary winding to burn out.

Theoretical background to the speed controller

Since the theory of operation of the induction motor is covered in numerous text books, it is sufficient here to concentrate on a basic equation that has relevance for speed control:

$$U/f = k\Phi$$

where U is the motor voltage, f the frequency of the motor voltage, k a constant, and Φ the magnetic flux of the stator field. The magnetic flux should, however, also be constant to prevent any likelihood of the tin-plate stator encapsulation running too hot as a result of saturation effects. Reduction of the flux, on the other hand, results in loss of torque. With this in mind, the above equation can be simplified to

$$U/f = \text{constant}$$

leading to the conclusion that frequency can be controlled as desired, along with a corresponding change in the motor voltage.

There are, of course, practical limits to such a simple equation. Notably at relatively low frequencies, when the ohmic resistance of the stator winding becomes significant, it is no longer allowed to change the supply voltage in proportion with frequency. This is mainly because increasing frequency means increasing

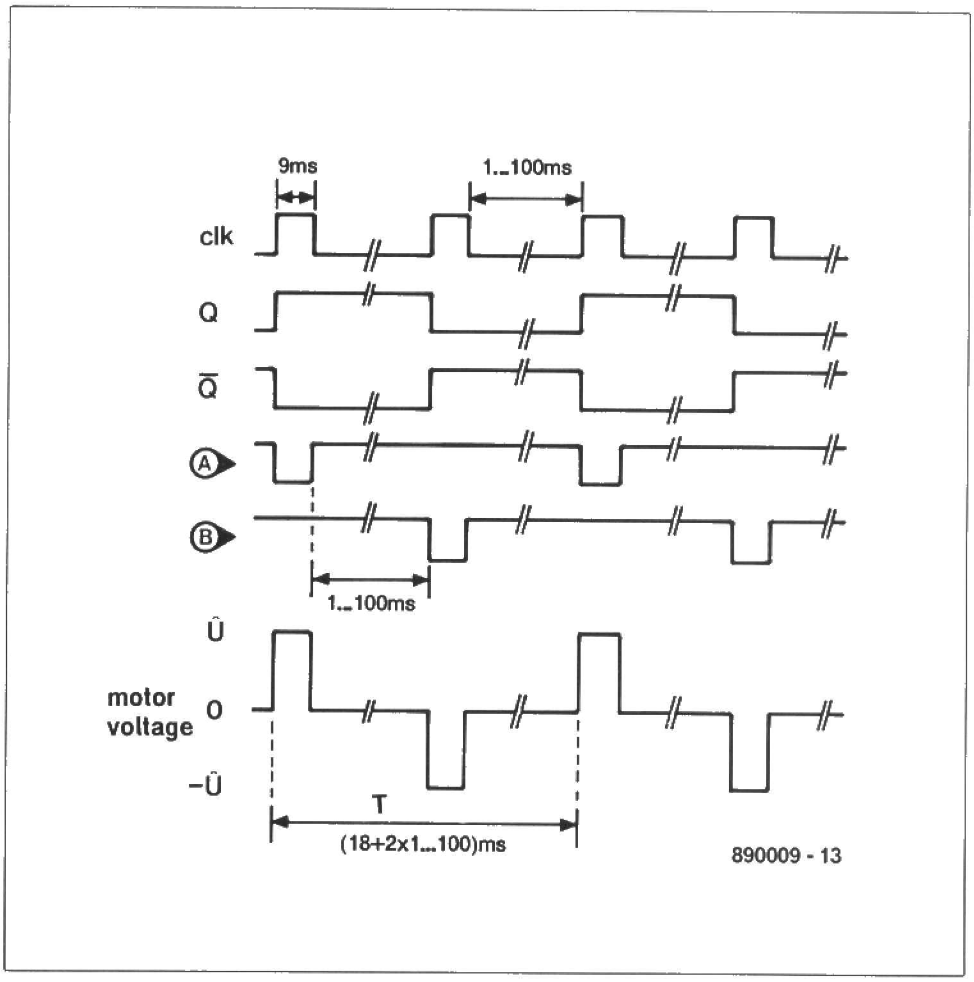


Fig. 3. Timing diagram showing the relation beween the main signals in the circuit.

the copper losses.
There is another factor that limits the use of the above equation: the breakdown voltage of the motor winding at a particular frequency. Fortunately, during normal operation of the motor, the safe limits for the applied voltage are not easily exceeded.

Design of a speed control circuit

The circuit diagram of the speed controller is given in Fig. 2. Although the circuit is offered as a design idea, it is none the less excellent for use with relatively small, lightly loaded, motors. The experimental nature of the circuit has mainly to do with the supply voltage and the way in which the motor is driven.
In the proposed circuit, the mains voltage is rectified by D_3 through D_6, which supply about $240\sqrt{2} \approx 339$ V. This voltage powers a half-bridge circuit that includes the motor. One disadvantage of the half-bridge is that the maximum motor supply voltage is limited to about half the rectified voltage. This results in some loss of torque, but has the advantage of avoiding damage to the motor when the ratio U/f is incorrect. Fortunately, there are ways to improve on the supply voltage, as will be detailed further on in this article.

The central part in the speed control circuit is multivibrator/bistable IC_1. The multivibator in this chip is configured as an astable multivibrator, whose frequency-determining network is connected to pins 1, 2 and 3. The charge and discharge time of C_1 can be controlled separately by virtue of D_1 and D_2. This arrangement results in a clock signal with a fixed 'high' time of 9 ms, and a variable 'low' time of 1 to 100 ms. The clock signal is divided by 2 in IC_1, so that the Q and $\overline{Q}$ outputs supply a symmetrical rectangular signal of frequency between 4.6 and about 50 Hz. Gates N_1 and N_2 combine the 3 output signals supplied by IC_1 into 2 pulse-like signals that are never 'low' simultaneously to prevent T_1 and T_2 conducting at the same time and causing a short-circuit. The pulses from N_1 and N_2 ensure, via T_3, T_4 and the primary winding of Tr_1, that an alternating magnetic field is created in the transformer. This field causes an alternating voltage of about 1 V_{pp} in the secondary. Rectification by the base-emitter junctions of T_1 and T_2, in combination with the anti-phase arrangement of the secondary windings, causes the transistors to conduct in alternating fashion.

The timing diagrams in Fig. 3 further illustrate the operation of the speed control circuit. The asymmetrical clock signal and the Q and $\overline{Q}$ pulses derived from it are shown in the upper 3 diagrams. Gates N_1 and N_2 generate active pulses A and B, which are 1 ms apart under all circumstances to guarantee that the output transistors are never on at the same time. The (ideal) motor voltage is drawn in the lower diagram. The peak value of it corresponds to half the supply voltage (about 170 V). The effective value, U_{rms}, is fairly simple to calculate from

$$U_{rms} = \hat{U}\,\frac{9}{(9+n)}$$

where $\hat{U}$ is the peak value, and n the period, which has a value between 1 and 100. It is seen that the effective voltage is inversely proportional to the period, and proportional to the frequency. The crux of the matter is, therefore, to find a combination of the fixed and variable part of the clock pulse, that resuls in a constant value of U/f. For a 240 V, 50 Hz, motor, this constant is $240/50 = 4.8$. In practice, however, this value is dimensioned higher at the lowest frequency. This is done chiefly to compensate the increasing influence of the ohmic resistance of the stator windings. A value lower than 4.8 is chosen at the highest frequency because it will be desired to regulate down to 50 Hz even with a too low supply voltage. The reduced torque caused by this compromise is not a problem in most cases.

Construction and safety precautions

The prototype of the speed controller was built on a small piece of veroboard. Since a part of the circuit is not insulated from the mains, all solder junctions are kept at least 6 mm apart, and non-used tracks are removed by overheating them with a high-power soldering iron. Also with safety in mind, the power section of the controller was built separate from the oscillator, in a moulded ABS enclosure.

Further considerations

The following is aimed at constructors wishing to experiment further with the circuit. It should be noted that the points raised below are mainly theoretical, and have not been put to the test in any practical circuit.
Capacitors C_2 and C_3 form the passive arm of the bridge circuit, and have a considerable influence on the motor voltage. The higher the motor power, the larger the capacitance of C_2 and C_3 to ensure the shape of the motor voltage. When these capacitors are too small, an oscilloscope will clearly show their being charged and discharged. For a 100 W motor, C_2 and C_3 should both be increased to about 25 μF.

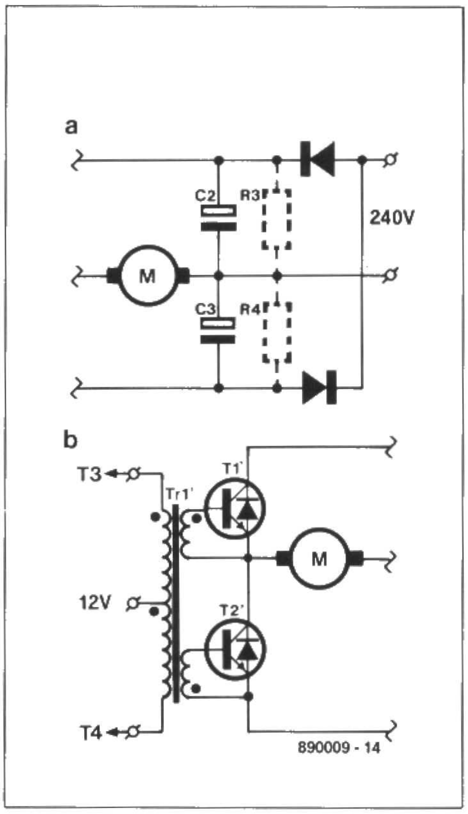


Fig. 4. Two alternative rectifier circuits to obtain a higher motor voltage.

Obviously, there is a clear point in raising the supply voltage of the bridge circuit. In principle, this voltage should be so high that the motor is fed with 240 V at 50 Hz. This supply level, U, is obtained from

$$U = 2 \times 240 \times \frac{9+1}{9} = 533 \text{ V}$$

This requires powering the circuit from $\frac{1}{2}\sqrt{2} \times 533 \approx 377$ V, which is an uncommon voltage. A voltage of 2×339 V is, however, could be made available by rectifying the mains voltage as shown in Fig. 4a. Unfortunately, this voltage is too high if the duty factor is not altered to maintain a correct ratio U/f.
An alternative rectifier, the full-wave type shown in Fig. 4b, ensures that the highest voltage remains below 339 V, but at the same time causes the peak value of the motor voltage to increase from 170 to 339 V.

In conclusion, designers should take due account of the duty factor and frequency of the control signal, as well as the supply voltage of the bridge circuit, since all these are tightly linked factors.

INTERMEDIATE PROJECT

A series of projects for the not-so-experienced constructor. Although each article will describe in detail the operation, use, construction and, where relevant, the underlying theory of the project, constructors will, none the less, require an elementary understanding and knowledge of electronic engineering. Each project in the series will be based on inexpensive and commonly available components.

2. Dark-room timer

This month's intermediate project is a low-cost timer that should appeal to the many photography enthusiasts convinced that the simple equipment in their dark-room can beat the print quality offered by high-street photo development centres. Providing an accuracy of one-third f-stop, the timer should be of particular interest for those who like to influence the contrast of the print by experimenting with different exposure times.

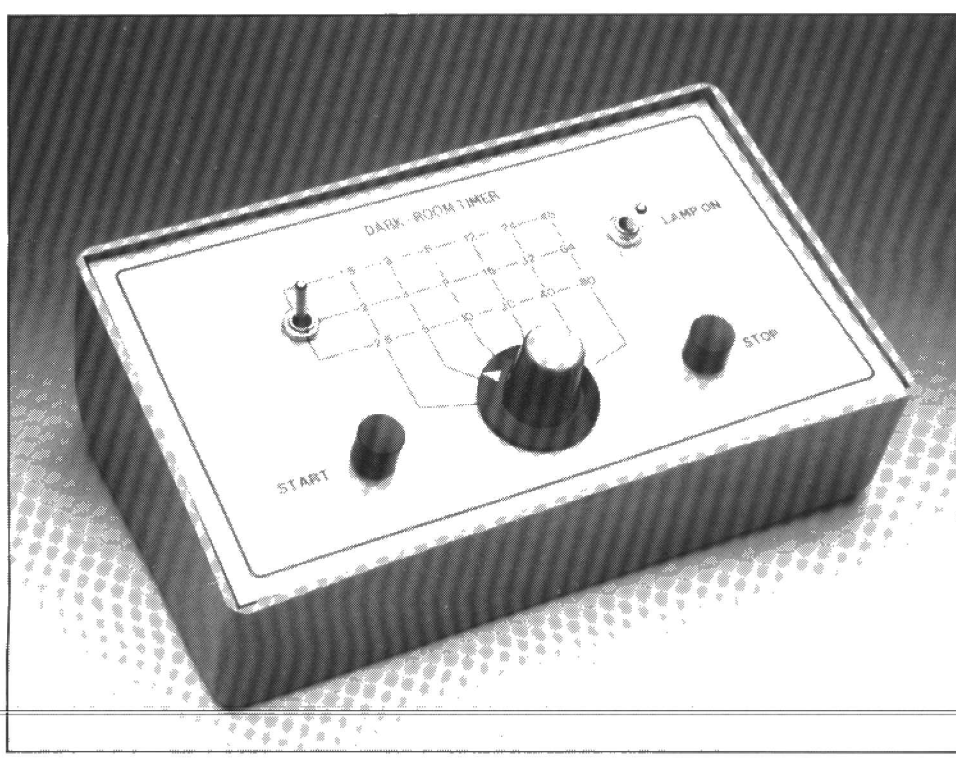
by A. Rigby

Optimum print quality of a photograph can only be achieved when the photographic sensitive paper is correctly exposed to light. In practice, this means that the photographer has to be able to accurately control the exposure time. With all the other, simultaneous, activities in the dark-room, such as handling the prints in the fixing and development baths, it is not so easy to devote all one's attention to timing the exposure. Many photographers would, therefore, like to use an electronic timer. These are available commercially, but the cost of even the simplest type may well be higher than the home-made type described here. This is so because the present timer is built from commonly available, inexpensive parts which may, for the greater part, be held in the junk box already.

The block diagram of the dark-room timer is given in Fig. 1. Rotary switch S_1 selects between three basic exposure times, which can be multiplied, or lengthened, by a factor 2, 4, 8, 16 or 32, as selected with another rotary switch, S_2. As shown in the above photograph, this results in 18 different exposure intervals in 1/3-f-stop increments, which is accurate enough for most applications. A continuously variable timer adjustment is purposely not used because this is cumbersome to calibrate and match to a scale that is accurate and at the same time easy to read. The function of the 'start' and 'stop' of the timer switches should be evident. Switch 'stop' may be pressed at any time to switch off the lamp — a very useful feature when it is found that an incorrect exposure time has ben set. A further switch, marked 'lamp on', is provided for adjusting the light source before the photosensitive paper is used.

For reasons of safety, the lamp in the enlarger is powered via a transistor-driven relay. To enable adjustment of the lamp, the driver transistor can be bypassed with the aid of switch S_5, so that the relay can be actuated manually also.

Bistable FF is the most essential part in the circuit. When it is set by pressing the 'start' switch, output Q goes high, so that the lamp lights until either 'stop' is pressed, or a reset pulse is generated by



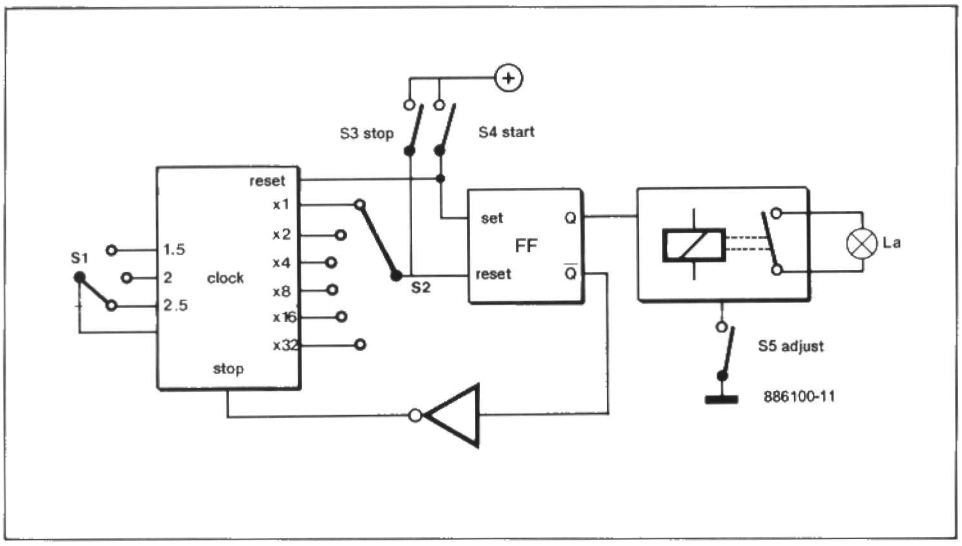


Fig. 1. Block diagram of the dark-room timer.

the clock oscillator. Since the set input of the bistable is connected to the reset input of the clock oscillator, this will be reset to state nought, and start counting down the lamp-on time, when 'start' is pressed. The lamp-on time is, obviously, determined by the positions of the rotary switches, which select the basic timing (S_1) and the multiplier (S_2) to give a total exposure time. Output $\overline{Q}$ (inverted Q) of the bistable is fed back, via an inverter, to the stop input of the counter/oscillator to prevent this counting on as the bistable has been reset. This arrangement is purely a matter of ensuring correct operation of the circuit at all times, since a single reset pulse is, in principle, enough for the bistable.

The circuit in detail

The circuit diagram of Fig. 2 shows that only the clock oscillator and the power supply are integrated circuits; the remainder of the functions are realized with the aid of discrete components.

The Type 4060, IC_2, contains an oscillator and 14 series-connected bistables (so-called *ripple cascade*). Internally, the oscillator signal is applied to the first bistable, which drives the second, and so on. Since each bistable divides its input signal by two, a total of 15 signals is, in principle, available, each of half the frequency of the previous one. On the 4060, 10 of these 15 signals are available on pins; 6 of them are used in the present application.

The parts connected to pins 9, 10 and 11 of the 4060, with the exception of R_1, determine the frequency of oscillation. Switch S_1 makes it possible to change the capacitance connected to pin 9, or the resistance to pin 10. The frequency of oscillation, f_0, is given by

$$f_0=1/(2.2RC).$$

With S_1 set to the centre position, factor *R* in the equation is formed by the sum of resistances P_1 and R_2. Factor *C* is formed by C_2, which has a value of 470 nF. If S_1 is set to the left position, the value of *R* is lowered since P_1 and P_2 are then connected in parallel. This means that the oscillator frequency increases. With S_1 set to the right position, C_1 and C_2 are connected in parallel, so that the oscillator frequency is lowered.

Since the IC forms a clock circuit, it may be more convenient to think in terms of periods rather than frequencies. The periodic time increases when S_1 is turned clock-wise (that is, in the circuit diagram).

With S_1 set to the centre position, f_0 will be about 8 Hz, which corresponds to a period of 0.125 s. This is not the minimum time increment, however, because dividers follow the oscillator. Pin 5 of the IC is output Q5, which supplies a signal with a period of 4 s (8 Hz divided by 2^5 equals 0.25 Hz), i.e., its high and low half-periods are 2 s long. It is this signal that determines the minimum time interval. When the 4060 is reset by pressing the 'start' button, all counter outputs are made low, and it takes 2 s before pin 5 goes high to reset the bistable.

Before the reset pulse generated by the 4060 can have any effect, the bistable must be set. This happens when the 'start' button, S_4, is actuated. A voltage is applied to the base of T_2, via R_6 and D_2. This voltage exceeds the base-emitter threshold, and causes the collector-emitter junction to conduct. This means that the collector voltage drops to about 0.1 V, which, in turn, results in T_1 and T_3 being turned off. The collector voltage of T_3 rises to about 5 V, so that T_5 is driven, and T_2 is kept conductive. T_4 ensures that relay driver T_5 can not conduct as yet. This is because T_4 receives base current via R_{11}, so that it short-circuits the base-emitter junction of T_5. This situation only ceases when the 'start' key is released.

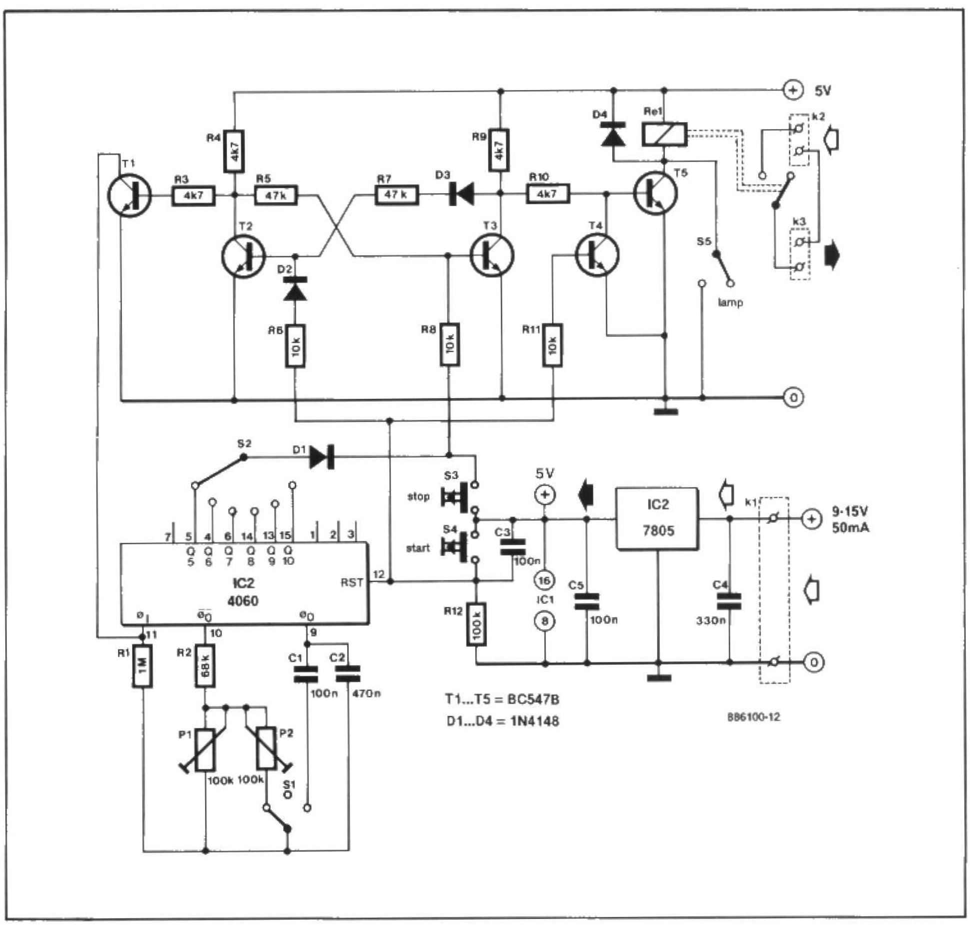


Fig. 2. Circuit diagram of the dark-room timer.

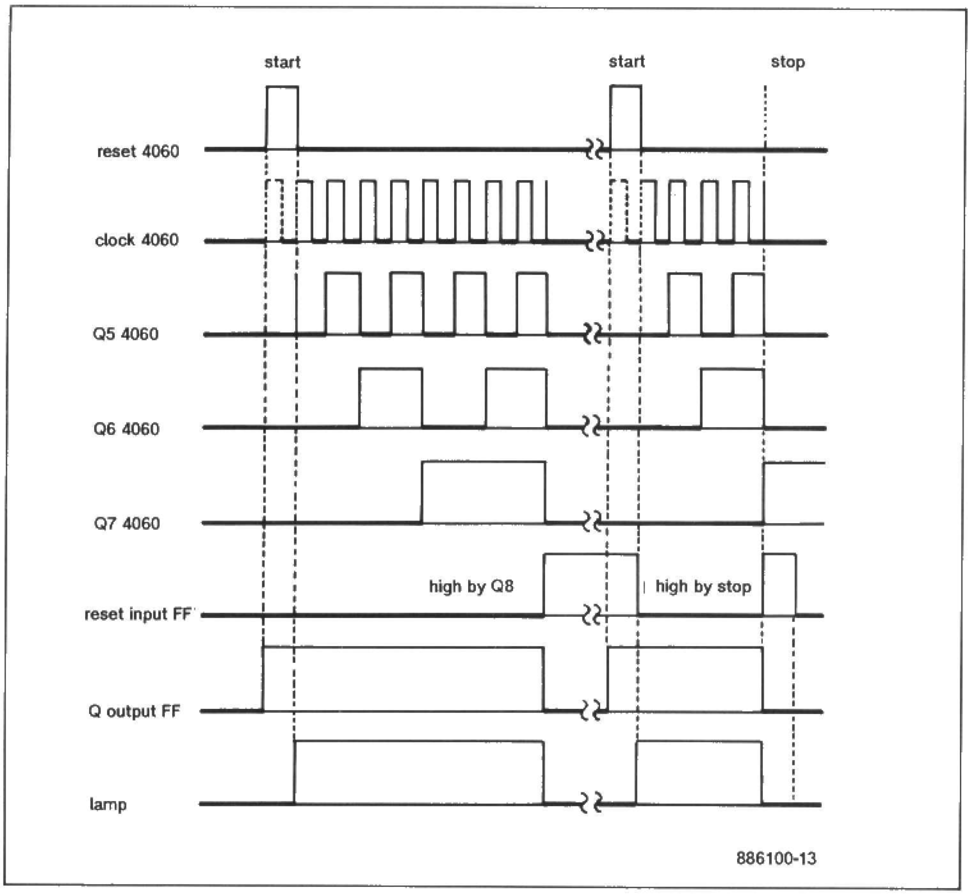


Fig. 3. Timing diagram. The dashed part of the clock signal indicates that the oscillator would be able to operate since T_1 does not conduct. This condition is, however, prevented by the level at the reset input.

Transistor T_4 is included in the circuit to ensure that the lamp lights as soon as the counter starts. Without the transistor, a timing error would arise, since the lamp would light when S_4 is pressed, and the counter would start when S_4 is released. When the selected time interval ends, or when 'stop' is pressed, the base of T_3 is taken high, so that the collector voltage drops. As a consequence, T_2 and T_5 are turned off, and the bistable is kept reset via R_5.

Timing diagram

The operation of the circuit is further explained in the timing diagram of Fig. 3. In this, it is assumed that S_2 is set to multiplication factor 8 (Q8). For the sake of clarity in the timing diagram, the frequency of the oscillator signal is shown lower than actual.

The first part of the diagram shows a normal timing event. A brief description of this is given below:

1. 'Start' key is pressed, but oscillator is not yet enabled by T_1 since reset line is still low; bistable output is high.
2. 'Start' key is released; oscillator starts; lamp lights, ripple dividers halve the oscillator frequency.
3. Output Q8 goes high; bistable is reset; oscillator stops; lamp goes out.

The second part of the diagram illustrates what happens when the 'stop' key is actuated after the circuit has been started. As expected, this reacts in the same way as it would have when a reset pulse is generated by the 4060.

Some more details

Diodes are fitted in three positions to obviate cross-effects between parts of the circuit. D_1 prevents the outputs of the 4060 being overloaded when they are low while the 'stop' key is being pressed. D_2 prevents the 4060 remaining reset when the collector voltage of T_3 is logic high. Finally, D_4 protects switching transistor T_5 against reverse EMF generated when the relay is de-actuated.

All resistors, with the exception of pull-down R_{12}, function as current limiters, while the decoupling capacitors in the circuit serve to eliminate interference.

Voltage regulator IC_2 allows the dark-room timer to be powered from an inexpensive mains adaptor with a DC output between 9 and 15 V.

All on one board

The printed-circuit board for the timer is shown in Fig. 4. Apart from the electrical components, the PCB also accomodates a number of mechanical parts, such as the relay, rotary switch S_2, and the two push-buttons, so that very little wiring is required. The direct supply voltage and the lamp voltage input and outputs are made in two-way terminal blocks for PCB mounting. Since most enlarger lamps work from the mains, great care should be taken in the connection of live wires to K_2 (mains input) and K_3 (lamp). The two Digitast switches ('start', S4, and 'stop', S_3) must be mounted on short pieces of relatively thick wire to enable them to protrude from the front panel, on to which S_2 is secured. Alternatively, S_3 and S_4 may be replaced by less expesive push-buttons for panel mounting.

Voltage regulator IC_2 can do without a heat-sink, and is bent back on to the PCB as shown on the overlay.

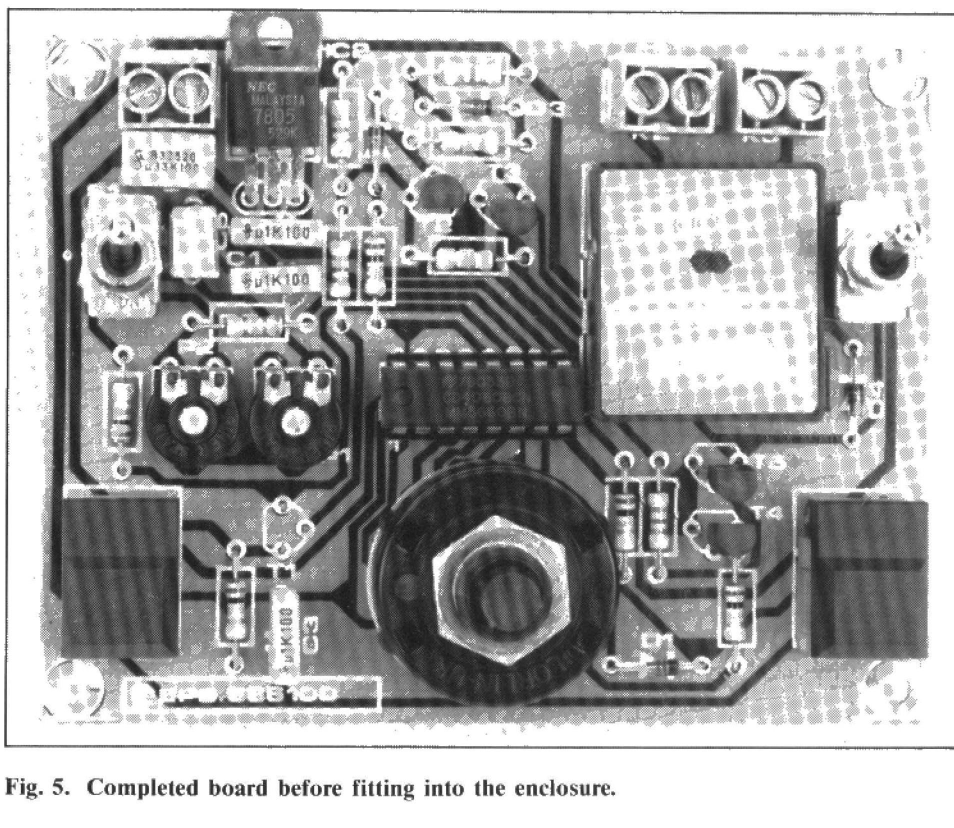


Fig. 5. Completed board before fitting into the enclosure.

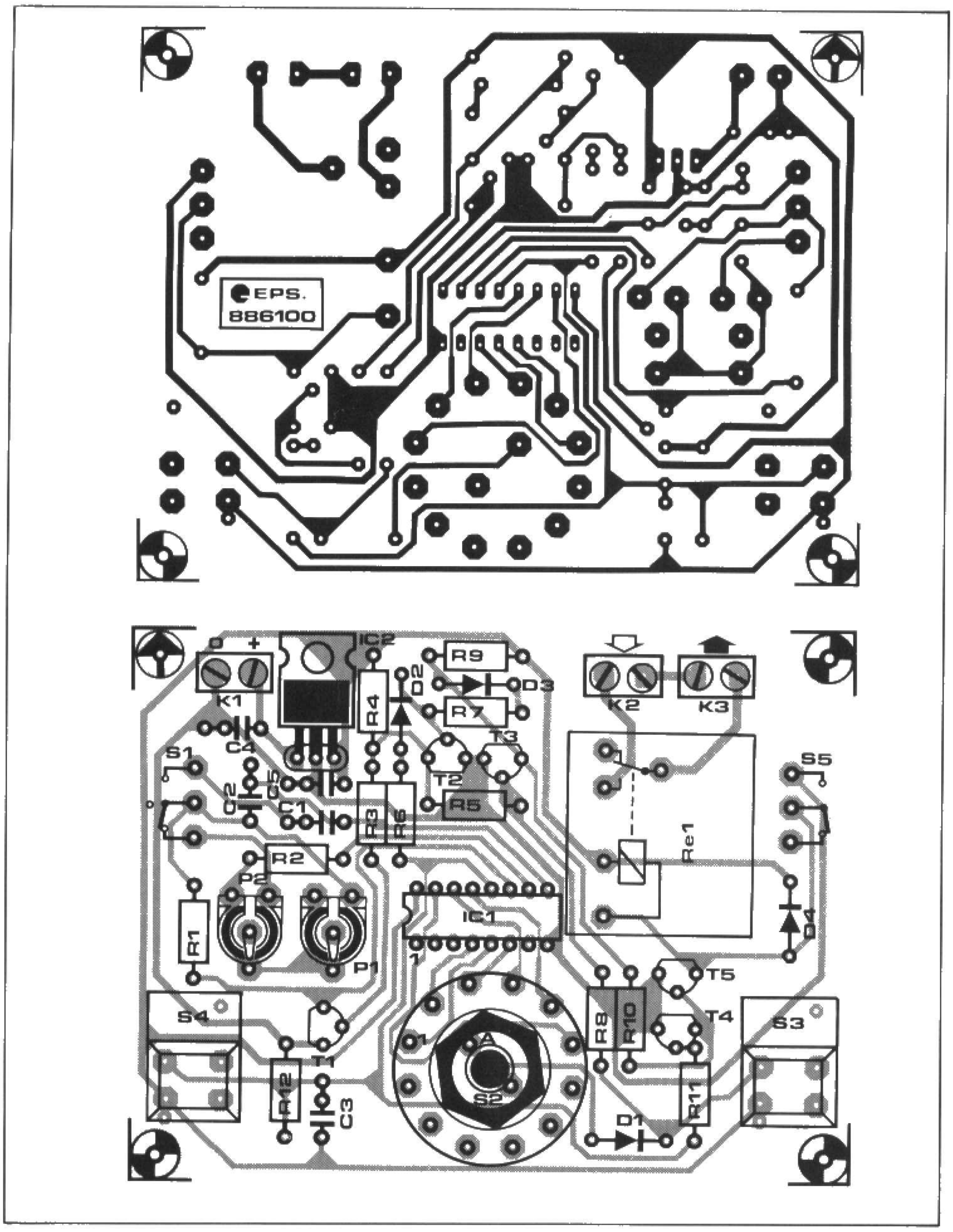


Fig. 4. Track layout and component mounting plan of the printed-circuit board.

Parts list

Resistors (± 5%):
R_1 = 1M0
R_2 = 68K
R_3;R_4;R_9;R_{10} = 4K7
R_5;R_7 = 47K
R_6;R_8;R_{11} = 10K
R_{12} = 100K
P_1;P_2 = 100K preset H

Capacitors:
C_1;C_3;C_5 = 100n
C_2 = 470n
C_4 = 330n

Semiconductors:
D_1 . . . D_4 incl. = 1N4148
T_1 . . . T_5 incl. = BC547B
IC_1 = 4060
IC_2 = 7805

Miscellaneous:
S_1 = miniature toggle switch with centre off position.
S_2 = single-pole 12-way rotary switch, or double-pole 6-way type.
S_3;S_4 = push-to-make button (see text).
Re_1 = V23127-A0001-A101 (Siemens).
K_1;K_2;K_3 = 2-way terminal block for PCB mounting.
Enclosure: e.g. Hammond Type 1595C.
PCB Type 886100 (not available ready-made through the Readers Services).

Setting up

The completed circuit should be set up carefully to ensure optimum results. First, set S_1 to the centre position, and S_2 in position '×1'. This setting corresponds to a timer interval of 2 s on the front panel. Then adjust P_1 until the relay is actuated for 2 s after pressing the 'start' key. This is a coarse adjustment, followed by advancing the rotary switch by 4 steps, and carefully re-adjusting P_1 for an interval of 32 s. The same is done for the 24 s interval to enable adjusting P_2. Finally, check whether the third position of S_2 yields the correct intervals. No adjustments should be needed at this stage, since this range derives its accuracy from the setting of P_1. If unacceptable deviations are observed, exchange C_1 by another type.

Finishing

The completed and aligned board (Fig. 5) is built into an ABS sloping-front cabinet as shown in the introductory photograph. As a finishing touch, provide the front panel with lettering and signs as suggested in Fig. 6.

Strain reliefs and good-quality grommets should be used where the mains cables enter the enclosure. In some cases, it may be preferred, however, to use a ready-made socket for the incoming mains voltage. It is imperative to ensure continuity of the earth line to the enlarger lamp when such a socket is used. ◼

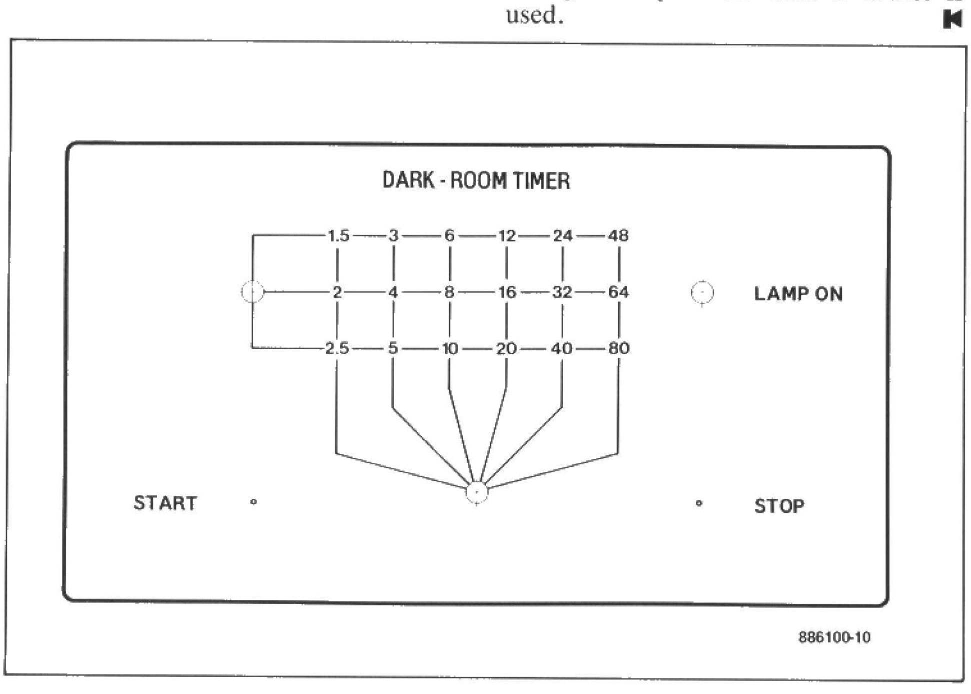


Fig. 6. Design of a possible front-panel for the dark-room timer.

COMPUTERS: AN OVERVIEW

by K.A. Roberts, BA

Early machines

It seems a far cry from the first automatic computer, the Automatic Sequence Controlled Calculator—ASCC. Yet, it is not quite half a century ago that this machine, the result of a collaboration between Dr Aiken of Harvard University and IBM, was presented to Harvard University in 1944.

Dr Aiken based much of his design on the Analytical Engine conceived in 1832 by Charles Babbage. Like Babbage's brainchild, the ASCC used sets of wheels as registers to store numbers. The machine was composed of no fewer than nearly 800,000 parts and almost 900 km of wire.

The first electronic computer

The first digital electronic computer came close on the heels of the ASCC: it was in full operation in 1946. Named ENIAC, acronym for Electronic Numerical Integrator and Calculator, it was designed by Dr J. Eckert and Dr J. Mauchly of the University of Pennsylvania.

Where the ASCC was a mechanical monster, the ENIAC was an electrical one: it contained some 18,000 electronic valves and consumed around 150 kW of electric power.

Not long after the ENIAC had been taken into operation at the University of Pennsylvania, a course of lectures was delivered at the same university that formed the mould for today's electronic computer. The lectures, *The Theory and Techniques of Electronic Digital Computers*, contained the principles for the design of electronic computers that had been worked out by a group of mathematicians and electrical engineers headed by a, now famous, Hungarian professor of mathematics working at Princeton, Johann von Neumann.

From then on, computer technology advanced at an accelerating pace. So much so that as early as 1956 Sir George Thomson, the eminent physicist, declared that

> 'the electronic computer has not made the headlines in the same way as nuclear energy, but I believe it is comparable in importance. The ability to apply precise reasoning to very large amounts of data in a reasonable time is something new, and the introduction of computers into science may prove not much less important than the introduction of mathematics in the seventeenth century'.

The mainframe era

During the 1950s and 1960s, the electronic computer evolved into a useful, but expensive tool. Scientific research, defence organizations, large accounts departments, and educational establishments had all begun to use some kind of computer.

The advent of time-sharing systems saw an even greater degree of computer penetration. The obvious advantages of allowing many users within a single company or organization to make simultaneous use of a central computer were quickly spotted by commercial entrepreneurs who set up commercial time-sharing services. Time sharing was for many the only way of making use of computer power: computers were still very expensive. Their cost lay not only in the initial outlay, but also in terms of operational staff, space and power requirements.

The minicomputer

Because of the high cost of computers, manufacturers realized the need for a smaller, relatively less complex (and thus less expensive) machines. The first manufacturer to bring one of these machines on the market was Digital (in 1963). In comparison with the mainframes then current, it was a limited machine: it ran only one program at a time, processed data in 12 bit-words and had only 4 k of memory. None the less, its advantages were obvious: it was not much larger than a domestic freezer, did not require an army of trained support staff, and its cost was only about 5—10% of the mainframe computers of the day when it was announced. Since it sold extremely well, penetrating not only new, but also existing, markets, its price rapidly came down, so that even more customers were attracted. By the early 1970s, no fewer than 70 US firms were manufacturing the so-called minicomputer.

Personal computers

Whereas the minicomputer came about for sound economic reasons, the microcomputer, perhaps better known as the personal computer, was, like radio in its early days, developed by amateurs. It should be noted that the industry at that time did not think a personal computer would ever cotton on (and this is only 15 years ago!). It was in 1974 that the July edition of *Radio Electronics*, an American hobbyist magazine, carried an article for the home construction of a small computer. The Mark 8, as it was called, used an Intel 8008 microprocessor, had 256 bytes of RAM (expandable up to 16 k) and had no ROM.

Despite its limitations, interest in the Mark 8 was phenomenal and sales of parts for it far exceeded expections. This interest, coupled with the introduction of Intel's 8080 microprocessor, prompted MITS, a small US electronics company, to introduce the Altair 8800. This design was also aimed at the hobbyist and designed for another American amateur publication, *Popular Electronics*. The project was published as a series of constructional articles, the first of which appeared in the January 1975 edition.

The computer was offered to readers of *Popular Electronics* for $650 fully assembled or $395 in kit form.

Apple Computers is born

Interest in the Altair 8800 caused the setting up, all over the USA, of 'computer clubs', run by, and for, amateur enthusiasts. A member of one such club in California, Stephen Wozniak, a self-taught computer engineer, got the idea of designing and manufacturing a similar kind of small computer, based on the newly-introduced 6502 microprocessor.

Wozniak designed a small computer, which was received enthusiastically by his fellow club members. However, when he approached his employers, Hewlett Packard, to try to interest them in manufacturing his computer, he met with a bland refusal. Hewlett-Packard did not think there was a sufficiently large market for the machine!

A friend of Wozniak's, Stephen Jobs, thought differently. He approached a number of potential buyers and eventu-

ally got a contract for a quantity of the Wozniak boards. Jobs and Wozniak thereupon went into business for themselves and formed what is now one of the largest computer companies in the world: Apple Computers. They have never looked back!
All this happened only 15 years ago. Today, the personal computer market far outstrips the mainframe and minicomputer markets, and tens of millions of PCs are in use the world over for a multitude of applications.

Parallel processing

It was stated earlier on that von Neumann's model formed the mould for today's computer. That was true until the arrival of the transputer. The processor in traditional computers can handle only a single instruction at a time. This is true even in multi-user and multi-tasking systems such as UNIX and concurrent MS-DOS, where the processor appears to be engaged in several tasks at a time, but in reality assigns time slots to portions of the relevant tasks. Obviously, the faster the processor, the less users are aware of the time-sharing process.
The transputer is a radical departure from the von Neumann concept. It is normalized for true concurrency. Parallel processing of data and instructions is achieved by synchronized very fast point-to-point communication channels between processes as well as individual transputer modules. There is, in principle, no limit on the number of transputer modules that can be connected to form a computer.
In contrast to other processors, transputers enable defining the speed of the system simply by adding as many modules as required. The IMS T800 transputer from Inmos, the designers and manufacturers of the transputer, in its 20 MHz version outperforms all of its 32-bit competitors, including the National Semiconductor NS32332-32081 and Motorola's MC68020-68881.
The calculation performance of the IMS T800 is equal to that of the VAX 8600 scientific computer from DEC, while a network of ten IMS T800 modules offers the speed and processing power of the Cyber 205 supercomputer from Control Data Corporation.
Because of their ability to work cooperatively in parallel on a number of different but related tasks, transputers are well suited for use in so-called parallel processing. By designing computers that work on a number of tasks simultaneously, instead of doing everything in sequence, designers aim to mimic more closely the workings of the human brain.
Transputers are also being assigned to less futuristic applications, including desk top supercomputers, laser printers and what have been nicknamed turbochargers where the transputer is used as an add-on unit to an existing system to upgrade its performance. High-performance graphics, engineering workstations, and robotics are other areas where the transputer has already begun to make an impact.

The optical computer

Beyond the transputer, research is going into photonic and molecular-based computers. Basically, the heart of a computer is the transistor (although there may be thousands of them on one IC). A transistor is nothing but a switch that can flip backwards and forwards between two states. Therefore, computers are chains of switches. They treat sequences of ons and offs to denote numbers (in which case ons and offs are read as the ones and zeros of binary counting) or to denote true or false (in which case chains of switches may be used as the building blocks of algebraic logic). Researchers at AT&T's Bell Laboratories and at Edinburgh's Heriot-Watt University have invented a device that does for light what the transistor does for electrons. This switch, known as a Bistable Optical Device—BOD—or transphasor, is essentially an optical transistor. Light emerges from it as a strong beam (on) or a weak one (off). Put a bunch of transphasors together, shine laser beams through them, and you have the basic ingredients of an optical computers.

The chemical computer

Even more advanced is the chemical computer that will operate in the same way as the human brain. The Science and Research Council—SERC—a few years ago set up a multi-million pound research project, called MERI—Molecular Electronics Research Initiative—that is intended to keep Britain in the forefront of advanced computer technology.
The idea of a molecular computer was first suggested by an American scientist, Forest Carter, as a means of overcoming heat dissipation problems in electronic computers. Living organisms are made up of carbon-based compounds, better known as organic compounds that inter-react to make possible, among many other things, such functions as thinking. Under the MERI, biologists and electronics experts will work side by side to engineer carbon-based chemicals that can replace electronic components now made from silicon. These chemicals will be able to interact at molecular level and will, therefore, provide enormous computing power in a very small space. Since molecules are interconnected in three dimensions, the computer based on them would be able to use parallel processing (like the transputer), making it very fast.
It would also be better at pattern recognition than conventional computers.

Some newcomers

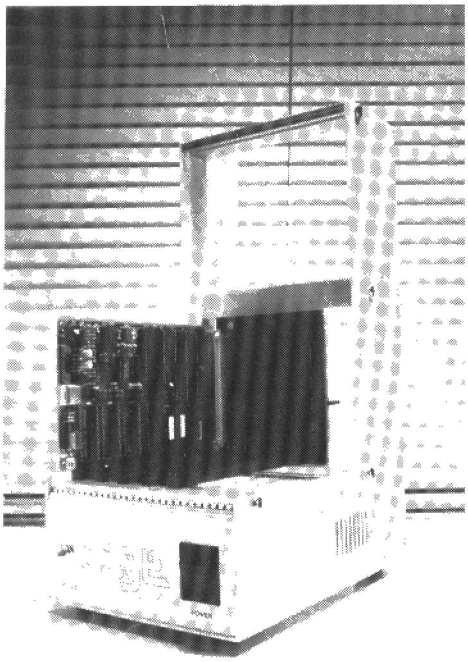
Back to today, one of the most exciting PCs to have come on the market in the past 18 months is undoubtedly the Archimedes. It is the first PC equipped with a 32-bit wide bus at a very reasonable price. Its processor is an Acorn Risc Machine—ARM—that is cheap and very fast. The high processing speed of 4 MIPS (million instructions per second) is the result of RISC (Reduced Instruction Set Computer) technology. The philosophy behind this technology is that it is better for the processor to work very fast from simple instructions than slowly from complex and often little-used instructions. Already, some versions of ARM have operated, under laboratory conditions, at processing speeds approaching 20 MIPS. It is noteworthy that although the ARM is comparable to Intel's 80836 chip in performance, its price is only about 1/100th of that of the 80836!
Another interesting introduction just over a year ago was from the man they can't keep down: Sir Clive Sinclair. His Z88 portable computer is cheap, small (smaller than a size A4 sheet of paper) and weighs just about 2 lb (less than 1 kg). All software is in ROM and it is not compatible with anything. The Z88 is intended as an end-product and comes, therefore, with all necessary software. Its ROM, apart from a number of tools, also contains a spreadsheet, a diary, a word processor and the well-known BBC BASIC. The programs may be used simultaneously. There is, of course, a serial connection for a printer so that texts from the word processor may be sent straight to the printer.

Finally

With all the kerfuffle about computers speeds, peripherals of a thousand kinds, software of unimaginable variety, it is sometimes well to reflect on the fact that a computer can really do only two things: carry out sequences of relatively unimportant operations like adding or copying, and choose between alternative sequences.

Entry-level STEbus system

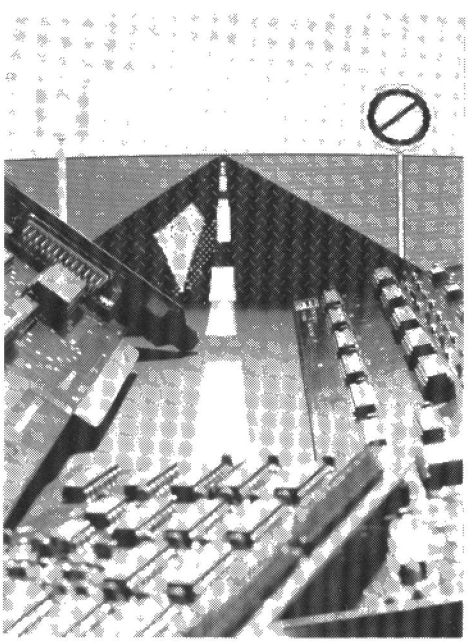
A low-cost development system for the STE backplane bus from Dage consists of a powered Fastframe enclosure with open sides for easy development access, and a single-board computer designed around the 8052 processor.
Apart from the 8052, this single-Eurocard microcomputer has four memory sockets, serial I/O, and an EPROM



programmer. The 8052's mask-programmed ROM contains an 8 K BASIC interpreter. This combination of facilities provides the basis for a low-cost STEbus development system, allowing the user to start program development simply by connecting a terminal with a standard RS232 interface. When this is done, the finished program can be blown into PROM in-situ for use in the final target system.

This 8052 STEbus system provides an efficient means of evaluating STEbus for industrial control applications. Priced at under £600, the system can be easily expanded over the STEbus and via VME.

Dage (GB) Ltd • Rabans Lane • AYLESBURY HP19 3RG



Low-cost quality IBM PC cards

Claimed to be 'probably the UK's lowest prices IBM-PC interface cards' are available from Flight Electronics. The range is extensive and falls into three main groups: general purpose, data communications, and industrial control. It includes a development board with built-in data and address buses and I/O line buffer; 8-, 12-, and 14-bit D-A and A-D converters; extensions; parallel/serial I/O; RAM expansion; monochrome graphic-printer; RS232-422; SDLC adaptor; 3270 BSC emulator; IEEE-488 (GPIB); and modem cards.

Flight Electronics Ltd • Flight House • Ascupart Street • SOUTHAMPTON SO1 1LU.

Portable PC from ITS

The ITS Portable PC has been designed to overcome what Integrated Technology Systems believe are the most common failings of existing portables.

The portable PC is IBM-AT compatible with a 12 MHz 80286 CPU, switchable to 6 MHz from the keyboard. Provision for an 80287 maths co-processor in-

dicates that the portable PC is suitable for engineers, designers, and others who need a high-performance portable they can use on site. 1 Mb RAM is standard on the main board and this can be upgraded to 4 Mb without using an expansion slot.

As well as from the mains, the PC can be powered from an internal battery that is automatically recharged when the PC is connected to the mains supply.

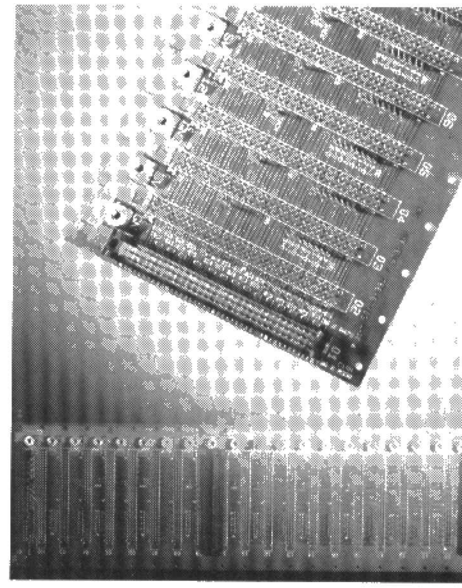
The high resolution backlit LCD display is easily readable in any lighting conditions and can handle software configured for standard displays, including EGA, where colours are represented by shades of grey.

Prices start at £2,250, excl. VAT, for the 20 Mb model.

Integrated Technology Systems • 5 Holyrood Avenue • Glenrothes • FIFE KY6 3PF.

The end of ground screen loops?

The VME-2000, a new design based on the industry-standard backplane VMEbus, has excellent cross-talk characteristics by virtue of an interference-free EMI-CAD (Elec-



troMagnetic Interference Computer-Aided Design) layout, negating the need of a ground screen loop. Cross-talk between adjacent signal lines, at 3.5 V steps with a 5ns switching time, is ±160 mV. That of traditional designs is typically around ±320 mV.

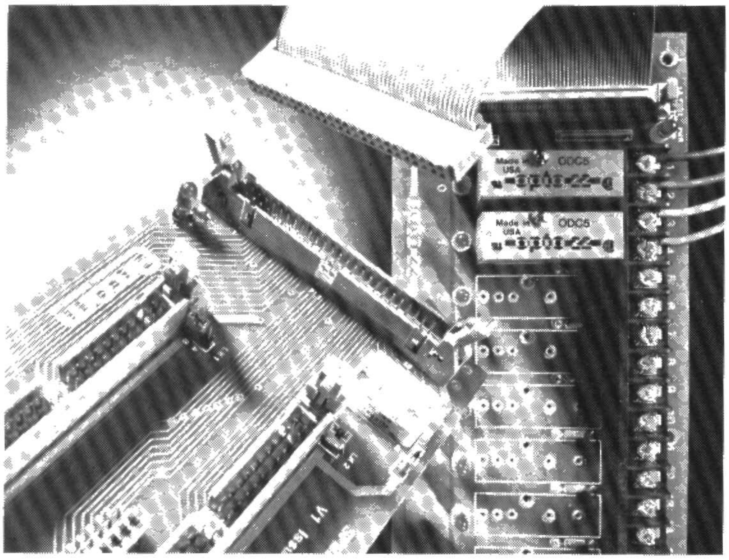
The system is based on the principle that a minimum number of layers reduces transmission signal problems, and fewer layers mean thicker dielectrics with better impedance characteristics. All lines change from one outer layer to the other at each connector position: consequently, they have the same length, and this ensures equal transmission times for all signals.

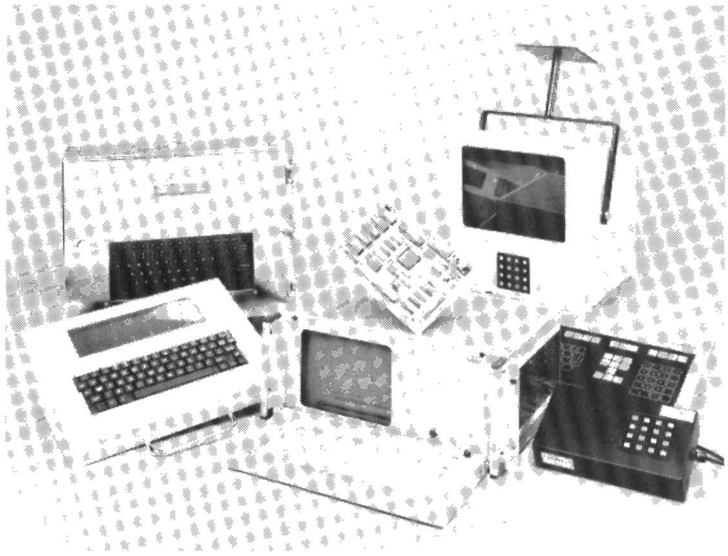
System Kontaks UK • The Paddock • Hambridge Road • NEWBURY RG14 5TQ.

Opto-22 interface for STEbus

The SPB22 interface board from Arcom provides connection between the standardized digital I/O of STEbus computer systems and the range of Opto-22 digital signal conditioning racks, widely used in industrial data acquisition and control systems. It enables the ungrading of engineers' computing requirements to world standard STEbus technology, while retaining existing plant wiring and the signal conditioning scheme of Opto-22 racks.

Digital I/O is taken direct from the STEbus computer board with a 50-way ribbon cable connection to the SPB22 interface, and this converts the standardized STEbus format to the Opti-ee scheme. The interface provides connec-





tors for up to two Opto-22 racks and these enable the STEbus computer to access up to 32 channels of I/O, again via low-cost ribbon cable.

Arcom Control Systems Ltd • Unit 8 • Clifton Road • CAMBRIDGE CB1 4WH.

User programmable terminal

The CAMDATA/188 is an industrial terminal that is user programmable from a standard PC. It may be thought of as a customized industrial PC at a fraction of the cost. It uses a 80188 microprocessor with memory to run a PC-DOS compatible operating system. Languages such as BASIC, Turbo Pascal, and C may be used.

CAMDATA Systems Ltd • The Old School • Church Street • SOMERSHAM PE17 3EG.

MAGNETIC DISKS AND TAPES

British Standard BS4783, in four parts, details how magnetic disks and tapes, whether used on a PC or on a mainframe, should be looked after.

Most people come into close contact with computers either in the home or at work. Some form of storage medium is required on which to distribute programs, to interchange data between computers, and to hold data off-line and when the computer is switched off. The most common type of storage medium in use today is some form of magnetic media. Even the smallest home micro uses either a magnetic tape cassette or a flexible disk cartridge. The larger computer installations often use either disk packs, disk cartridges or magnetic tape on open spools, as well as flexible disk cartridges and magnetic tape cartridges.

BS 4783 has been prepared under the direction of the Information Systems Technology Standards Committee. It has been developed from advice and information provided by Government computing departments, magnetic media suppliers, manufacturers and users. Magnetic media are manufactured from carefully developed high grade materials. Modern production techniques, followed by rigorous testing and inspection, ensure that the products supplied to the user conform to exacting standards and are capable of a long life. To continue to benefit from the care taken during manufacture, and to ensure optimum performance from a product during its life, the user should exercise care; this British Standard gives appropriate advice.

The larger computer installations have air-conditioning, full-time operators, a media library, on-site engineers and a workshop with test and media cleaning equipment. Careful records are kept of media, with dates received, usage and errors. Secondary copies of important data are maintained, possibly at a different site.

At the other end of the scale is the computer used at home or in the office. Here there is no air-conditioning, media may be stored in drawers in the same area as the computer, operators have other responsibilities, no facilities are available for cleaning media and there is no established procedure for maintaining copies.

These documents give recommended practices that should be adopted in full at larger installations, where facilities are available and where the disruption caused by breakdown due to media failure would be serious. At smaller installations users are advised to adopt as many of the recommendations as possible, balancing the cost of implementation against the cost of media failure.

The publications issued by manufacturers of magnetic media, and of equipment for handling it, also give advice; they should be read in conjunction with this standard.

Though each of the four types of magnetic media discussed in the standard has specific requirements, many recommendations are common to all four Parts of the standard.

Media should be stored in a library, administered by a librarian with responsibility for preparing and maintaining a history record for each item. The library, operating area and equipment must be kept scrupulously clean. Advice is given on discipline to achieve this. In spite of all precautions media will require cleaning either on a routine basis or when errors occur. Routines, and suitable equipment and materials are discussed.

The physical properties of media will only be permanently affected by extreme temperatures and humidities, but reliable and consistent operation require that the media be used, stored and transported within specified ranges.

Media require visible labels; these would appear to be simple to acquire, mark and apply. However, the adhesive must not creep, and writing on a label which has already been applied may damage the media.

BS4783 is being issued in four Parts as follows:

Part 1 Disk packs, storage modules and disk cartridges.

These products are based on rigid aluminium platters coated with a magnetic surface. In operation they rotate at high speed; the write/read heads fly above the magnetic surface, floating on the cushion of air induced by the high surface speed. As the separation can be as low as 0.4×10 m it is essential that high levels of cleanliness are maintained and that the disk is not distorted. Dust, debris, or even grease on the surface can cause errors due to instability of the flying attitude of the head, and can, in extreme cases, result in the head landing on the surface and scoring it. The recommendations on storage, handling, transportation, inspection, cleaning and maintenance are directed at cleanliness and the avoidance, or detection of, damage to the disk. Advice is also given on the identification of faults.

Part 2 Magnetic tape on open spools.

This Part includes recommendations for storage, handling, transportation, inspection, cleaning, maintenance, fault identification and fault recovery. The photographs which are included give examples of tape conditions that may lead to poor performance.

Magnetic tape runs in contact with the write and read heads; on some tape transports tape guidance is over air bearings, on others the guides are in contact with the tape; the production of some debris in the transport is inevitable. Attention must be given to the cleanliness of the transport as well as of the tape as a transport with dust or debris on heads or guides can cause damage to a series of tape.

As magnetic tape on open spools is the preferred medium for archives, procedures for such long term storage are described.

Part 3 Flexible disk cartridges

Flexible disk cartridges are typically used, and in large numbers, in desk top computers, terminals and test equipment. There is a danger that the very prevalence of these cartridges, and their apparent simplicity, will obscure the fact that their reliable operation nevertheless depends on careful treatment. This Part includes recommendations for storage, handling, transportation, inspection, maintenance and fault identification. Illustrations are included of damaged cartridges.

A flexible disk is encased in a protective jacket that has windows to enable the write/read heads of the drive to access the surfaces. The write/read heads are in contact with the surfaces but in this product a liner inside the jacket absorbs the wear products as the disk rotates. The greater danger to reliable operation is physical damage suffered if the cartridge is allowed to lie on a desk or table, where it may become dented or deformed, or the disk may be contaminated by finger marks through the head windows.

Part 4 Magnetic tape cartridges and cassettes.

Here the tape is protected from physical damage whilst handling by being totally enclosed except for the aperture required to enable the head to access the recording surface. The tape runs in contact with the write/read head but surface wear is reduced by the use of few tape guides. Care must be taken to prevent the case becoming chipped when loading or handling as the debris may appear in the tape path. This Part includes recommendations for storage, handling, transportation, inspection, maintenance, and fault identification and recovery.

BS 4783 is available either in individual Parts, or in a package of all four Parts from BSI Sales, Linford Wood, Milton Keynes MK14 6LE. Price: Parts 1, 2 and 3 £28.50 each (£14.25 each to BSI subscribing members), Part 4 £17.00 (£8.50 to BSI subscribing members).

MOLECULAR ELECTRONICS

Towards an advanced form of computer technology

by John Delin

Smell sensors, paper-thin television screens, moving holograms and bio-computers using living organisms are all the stuff of fantasy. But so was the silicon chip or the seemingly incredible notion that a million pieces of information could be stored in a grain of dust, only a few decades ago.

Today's microprocessors are hardly obsolete but they are already revealing limitations in terms of size and other physical constraints. Where then can scientists turn to complement existing technologies while providing an exciting springboard into the fantasies and realities of the future?

Molecular electronics, designed to harness the molecule itself as an information processor, show considerable promise. Gathered together from disparate research over a range of disciplines, this line of thought has attracted considerable interest in Britain and has now been selected as one of the major areas of the Department of Trade and Industry's new Link[1] programme of collaborative research between universities and industry.

At least £20 million is to be allocated to molecular electronics, half from government sources and half from industry, to cover the so-called pre-competitive stage of development, delving into fundamental principles and the feasibility of devices. The programme aims to provide the platform from which industry and industrially sponsored research can later develop exploitable products.

Practical application

Molecular electronics uses organic molecules to process information. It goes beyond the digital processing of conventional electronics and adds new dimensions—for example structures and shapes—to its vocabulary. Conventional electronics are analagous with the nerves in the body that trigger when a certain electrical threshold is reached. Molecular electronics resemble the white corpuscles that react to the shape, density or temperature of a bacterial invader. They are conceptual in action rather than computational.

One familiar example of molecular electronics in action is the liquid crystal display seen in watches and calculators that respond vigorously to electrical or heat signals. These are already in use in a number of

applications such as the head-up displays where an array of indicators is visible in the windscreen to aircraft pilots or railway locomotive drivers. The operator is not distracted by having to search for gauges but can absorb vital information such as engine speed or temperature without moving his head.

Unfortunately today's liquid crystals are sluggish in action and not very sensitive. The race is on to improve them and the newer ferro-electric crystals already being tested are more flexible and responsive. Displays are now envisaged that will reproduce four-colour signals on a flat screen — an obvious precursor to wallpaper television.

Cheap mass memories

Using the ultra-thin films of one molecule thickness now being produced, coupled with photo-electronics, one could envisage quickly changing holograms, no more than a short step conceptually from genuinely three dimensional moving pictures.

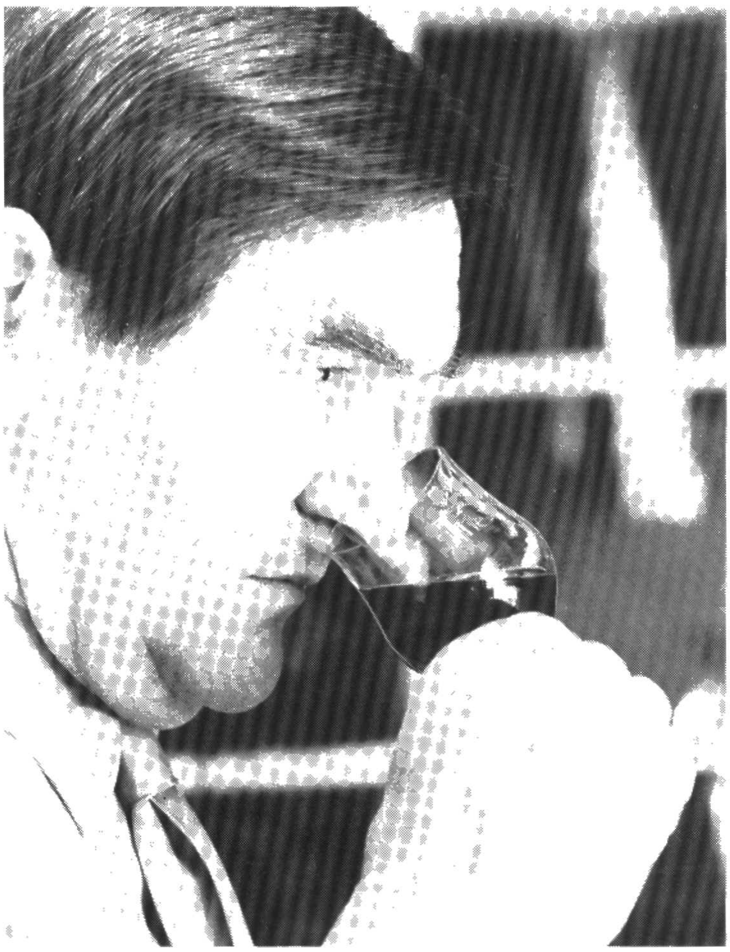
Britain's Link programme defines molecular electronics as "systematic exploitation of molecular, including macromolecular, materials in electronics and related areas such as photo-electronics". Liquid crystalline substances apart, it proposes to investigate organic metals and semiconductors, non-linear optical materials, and photochromic, electrochromic, piezo-electric and pyro-electric substances. Applications to be studied include information storage and transmission, signal processing and thin film technology.

Experts see the envisaged molecular electronic systems as compact, flexible, cheap and efficient. According to Professor John Barker of Glasgow University, a leading worker in the field, the smallness of molecules make possible very dense circuits leading to cheap mass memories.

Molecules are much more noise stable and thermally stable than conventional semiconductors. Metallic, semiconducting, insulating molecular components to be built into switches and circuits would be the very least one might expect.

Artificial intelligence

Compared with the paucity of good solid-state electronic materials, molecular materials have a rich variety of possibilities ranging from simple small molecules, polymers and molecular crystals to complex macromolecules bordering on biologically significant structures not far removed from natural organisms. Many of the techniques employed in synthetic organic chemistry could provide the means of building useful additional properties into such materials.



Will an "electronic nose" ever replace the wine expert's experience and discrimination?

A significant example of the last technique is illustrated in the electronic nose project being conducted jointly at Glasgow University and Warwick University. By varying the structures of a range of polymers it has been possible to devise a machine that can smell. It converts the responses of a range of sensors into signals related to specific smells.

In modern sciene it has apparently been the most simple senses such as taste, smell and sight, particularly visual recognition, that have been most difficult to reproduce in their true complexity. Such advances as these will offer important insights into the way the brain works and the development of artificial intelligence.

Many scientists see the way ahead as grafting molecular electronics on to the conventional variety, possibly using thin films as an interface and producing, in effect, conventional microprocessors with a much wider range of sensitivities.

Insect colony

For example, today's industrial robots are firmly based on the digital principle, essentially by measurement and counting. The concept of a robot that could compare, sense and feel brings in an almost occult dimension.

The logical conclusion to this development, as some scientists see it, will be to bridge the gap between molecular and biological structures, linking the animate and inanimate and producing living intelligences—the bio-computers of the future. It does not follow that these will be human brain analogs.

In fact, part of the excitement will be the construction of different types of intelligence to supplement the human variety. We are already familiar with such variations, as for example, the collective intelligence of an insect colony as compared with the more isolated individuality of the human brain.

These distinctions are already accepted by most scientists working in molecular electronics. The theory of molecular computing differs radically from the now conventional digital computing which, strictly speaking, is no more than advanced counting. On paper at least, researchers are already producing new structures and simulations that presage many fresh approaches to artificial and natural intelligence.

Hope and prejudice

The universities and industrial groups collaborating in the Link molecular electronics programme aim to produce exploitable hardware within the next few years and expect to achieve consistent rather than dramatic development within this time. Professor David Bloor of Queen Mary College[2], London, who is the Link programme coordinator for molecular electronics, is quite clear on the point.

"Molecular electronics is not a replacement for the silicon chip", he said. "We are prejudiced by silicon technology and people must be encouraged to take different attitudes. Molecular electronics, if it comes to fruition, will be as different from today's electronics as semiconductors are from the valve technology of 40 years ago."

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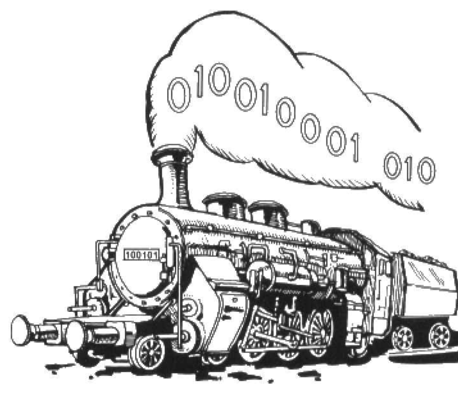
(1) Link Secretariat, Department of Trade and Industry, Room 237, Ashdown House, 123 Victoria Street, London SW1E 6RB.

(2) Professor David Bloor, Department of Physics, Queen Mary College, Mile End Road, London E1 4NS.

THE DIGITAL MODEL TRAIN — PART 1

by T. Wigmore



As every railway modeller knows, the control of model railways is being transferred inexorably from the heavy-duty switches and relays of yesteryear to the digital computer. In a new series of articles, we describe a number of units based on the new technology, culminating in a fully electronic model railway. The series commences with a description of the Marklin control system in which all commands to the signals, locomotive and points (turnouts) are given via the rails.

Although in this and some future articles reference will be made to the Marklin system, it should be noted that a number of units will be described that may not only replace the relevant Marklin circuit, but can be used in a variety of DC railways of other manufactures.

Rails: a serial bus

In any model railway, there are a number of operations that must be under full control at all times. Points (turnouts) and signals may be operated independently of one another in a simple manner, because they all have their own power and control connections. The drawback of this type of parallel control is the ensuing complexity of the wiring. It is far more complicated to control locomotives independently, because their only contact with the "driver" is via the rails. There are control systems that provide a number of high-frequency command signals. Each locomotive is then fitted with a special filter that allows it to be operated on one specific frequency only. Even these systems are limited to 10 or 15 independent locomotives, because the operating frequencies must be spaced fairly widely to ensure complete freedom from interference. However, time has already caught up with these systems.

The Marklin system is unambiguously based on computer technology. It makes use of a two-wire bus (communication channel) that is already present in any model railway: the rails. Each item to be controlled is connected to the rails (from which it is also powered) and given an address. When a given item, be it signal, locomotive or point (turnout), is to be operated, the relevant address is entered on to the bus followed by a data stream that contains the operating command. It is clear that each item needs an address decoder that will indicate when it is being addressed. The data stream contains a certain measure of redundancy to obviate erroneous operations. This is particularly useful with locomotives, because the frequently bad contact between wheels and rails is a real source of trouble.

The command signal is entered on to the rails by the central control computer in packets of nine bits (strictly speaking, the supply voltage is being modulated). Of the nine bits, the first four (in locomotive decoders) or five (in point—turnout—decoders) are accepted as address bits and the remainder as data bits. It is noteworthy that the so-called trinary system is used for the address bits. In this system, a bit can have three states: logic 0, logic open, and logic 1. The protocol of these states is shown in serial format in Fig. 2.

It is because of these three possible states that a fairly large number of addresses

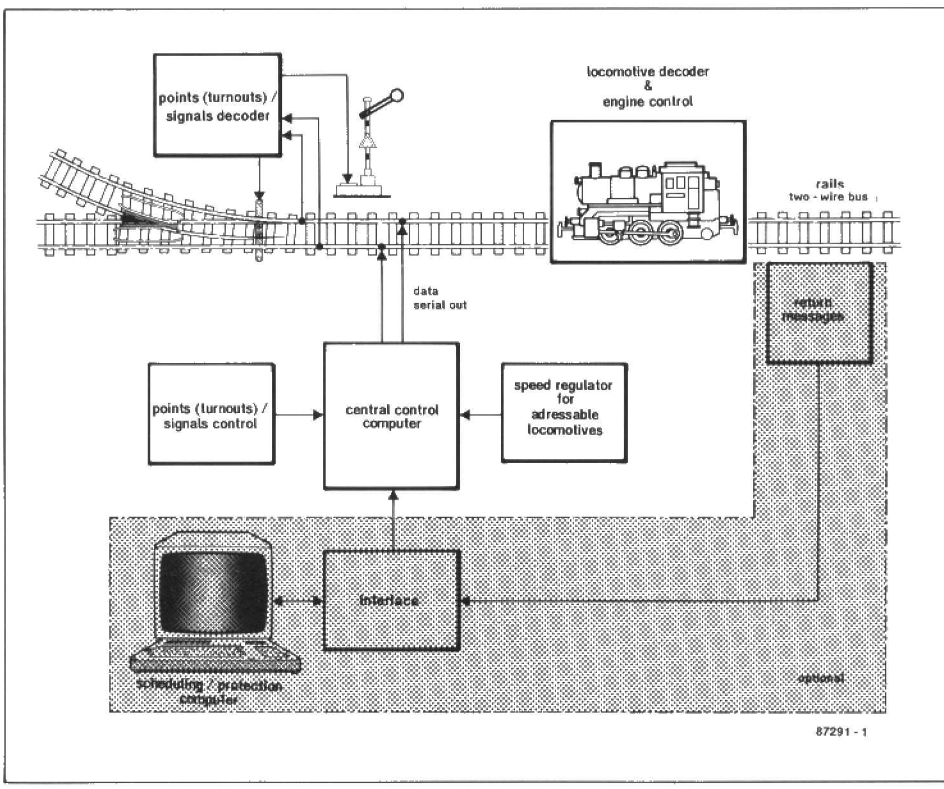


Fig. 1. Block schematic of a digital model railway as designed by Marklin. The rails are used as a two-wire bus.

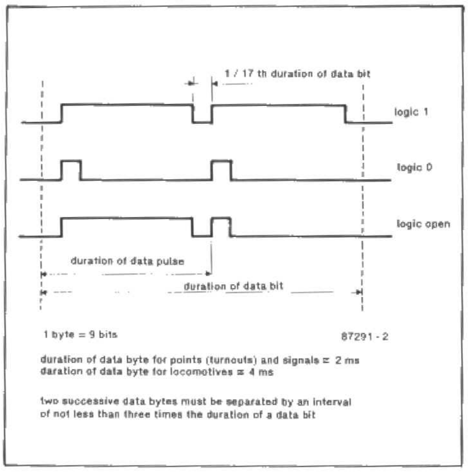


Fig. 2. The serial data format. The short pulses serve as markers. The three states of the trinary system are: 00 = logic 0; 10 = logic indeterminate; 11 = logic 1.

may be obtained with relatively few bits (i.e. connections): 81 for locomotives and (theoretically) 243 for points (turnouts).

Furthermore, the baud rate of the locomotive control signal is half that of the signals and point control signal. Apart from ensuring a more reliable data transmission to the locomotives, the use of different baud rates enables extending the address range. It should be noted that the decoders for locomotives and points (turnouts) operate in the same address range (except, of course, bit 5 which is a data bit for locomotives and an address bit for points). The decoder merely ignores signals with a baud rate different from that for which it is designed.

A practical circuit: point/signals decoder

We have chosen a relatively simple circuit to describe the Marklin system. The decoder in Fig. 4 may be used for the control of up to four points (turnouts) or signals. The serial data extracted from the supply voltage via R_7 and the clamping diodes on board IC_1 are decoded by IC_1. The first five bits are accepted as address bits. However, input A_5 is connected to ground, so that only one third of the address range is reserved for the points and signals, i.e., theoretically, 81 decoders may be connected. Each decoder is given a trinary address with the aid of shorting plugs or wire bridges (see also Table 1.).

The total number of points (turnouts) is restricted to 256, because not more than 16 switching boxes (each with switches for 16 points) can be connected to the central computer. Evidently, not all trinary addresses are used.

In each decoder, three of the four data bits are used to form a sort of sub-address that serves to select one of eight possible magnet coils. This is done with the aid of a 3-to-8 decoder, IC_2, which, on the command of the last data bit, connects one of the darlington inputs to the positive supply line via R_2. In the circuit, use is made of the darlingtons contained in a ULN2001A, because this device is relatively cheap. It also contains

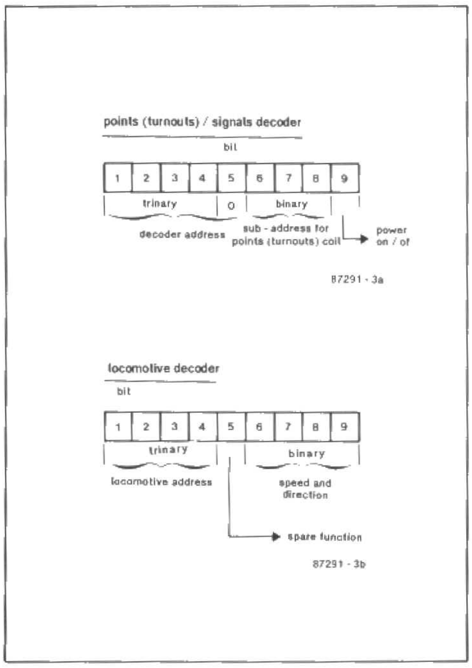


Fig. 3. Construction of the 9-bit data words that control points (turnouts) and signals (a) and locomotives (b). The address bits operate with three logic states.

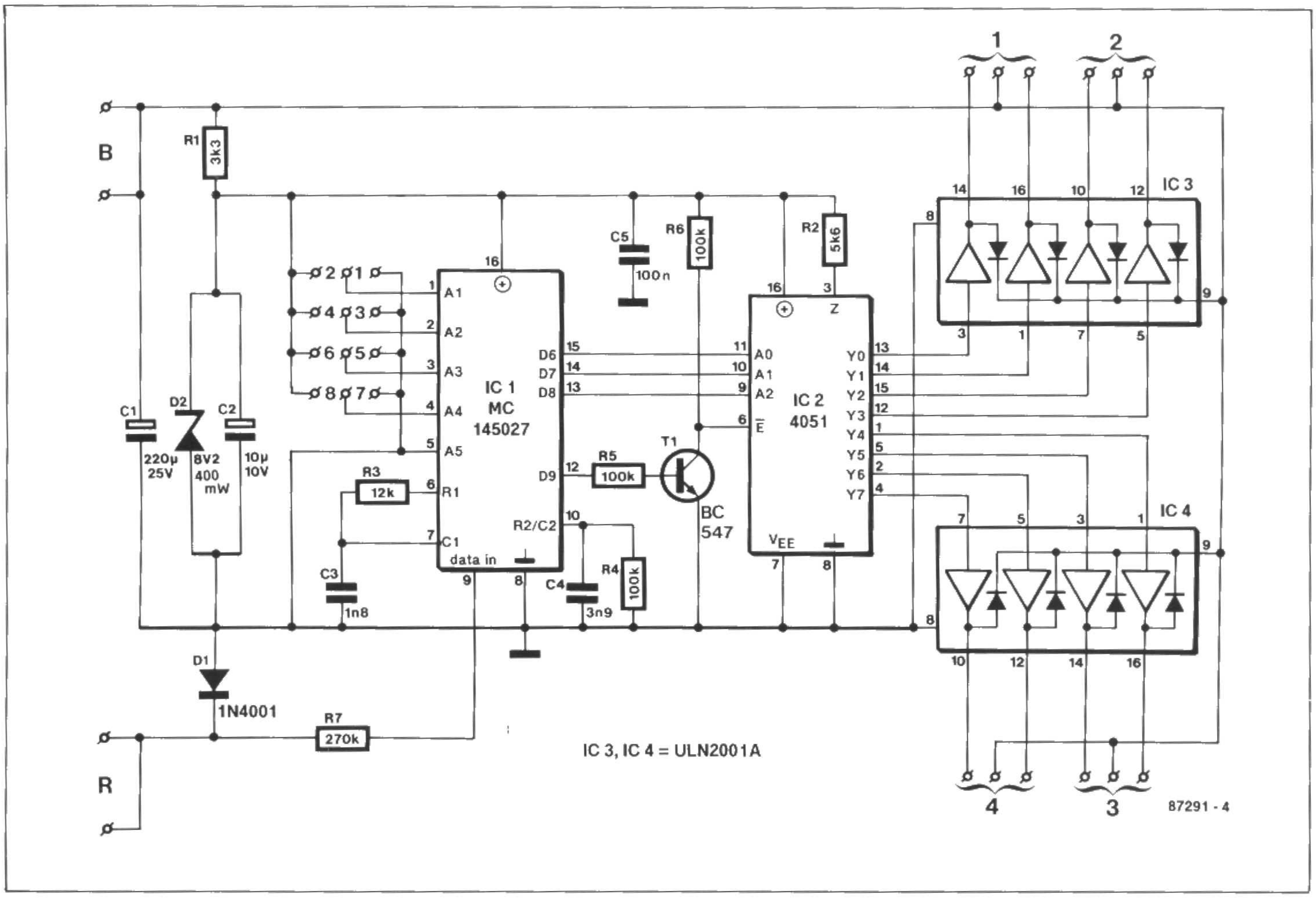


Fig. 4. Circuit diagram of the decoder for points (turnouts) and signals.

a number of indispensable freewheeling diodes. These diodes prevent the high voltage peaks generated by the on and off switching of inductive loads being superimposed on the supply voltage.
A little more detail about IC_1: see Fig. 5. Network R_3-C_3 is used to differentiate between short and long received pulses. The short ones may be considered as "markers"; the trinary information is contained in the intervening long pulses. Time constant R_4-C_4 serves to separate sequential data words.
If the received address, i.e., the first five bits of a data byte, matches its wired-in address, the decoder will transfer the received data to a 4-bit shift register. They are not yet available at the outputs. Only when a second, identical, data word is received are the data transferred to the output register. This arrangement ensures a large degree of freedom from interference.

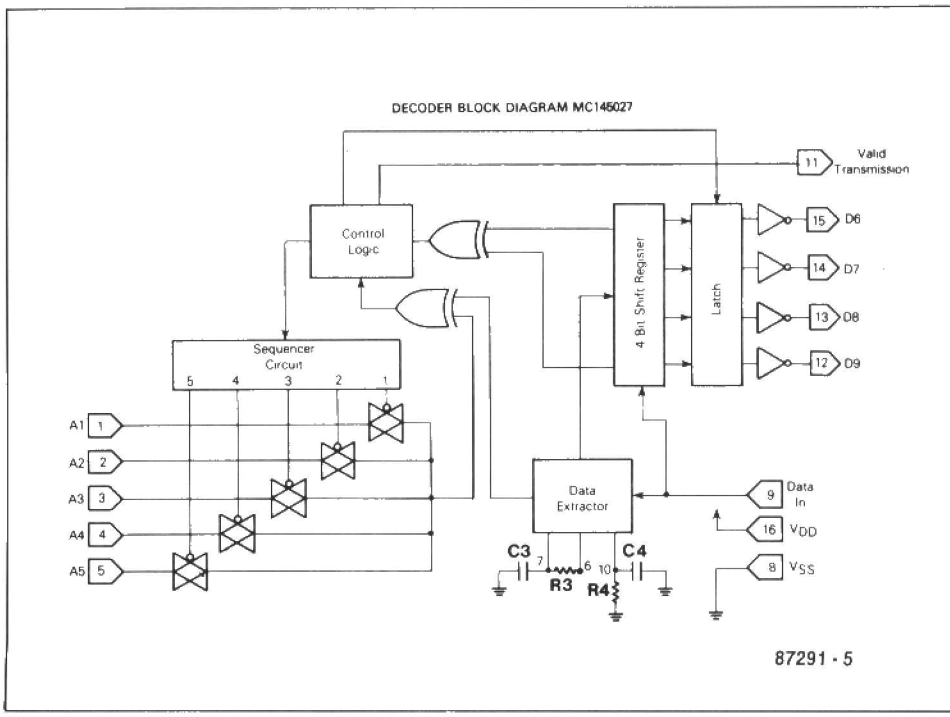


Fig. 5. Block diagram of the MC145027.

Price/performance considerations

It may not be clear what the advantages are in using points (turnouts) decoders instead of conventional wiring and relays. After all, the saving in wire does not compare with the cost of a decoder. The main advantage of a decoder is that it affords the possibility of "intelligent" control of points (turnouts). The "intelligence" may take the form of pre-programmed switching of combinations of points (turnouts) or of computer-controlled scheduling and protection.
It is, of course, not possible to power each and every locomotive via separate wires. The advantage of a decoder is here, therefore, much clearer.

In principle, a locomotive decoder works in a similar fashion as that for signals and points (turnouts). Four bits are used to address a locomotive (up to 81 may be used). Bit 5 is used for special functions and the remaining four bits serve to control speed and direction.
Since locomotive power is present on the rails at all times, permanent train lighting presents no problems. It would be feasible to power other aspects, such as station lights, via the rails, but in view of the maximum current the central unit can provide, it is wise to power equipment not directly connected with the rolling stock from a separate supply.

The practical side

Constructing the points (turnouts)/ signals decoder on the printed-circuit board shown in Fig. 6 should not present any difficulties. Connecting it to the track is no problem either. There are two connections: red and brown and these are connected to the corresponding terminals of the Marklin system. Table 1 shows how the short-circuiting jump wires are to be located for setting the various addresses.
Each point (turnout) or signal has three terminals. The central one of these is used for the common wire of the two

Parts list

Resistors (±5%):
R_1 = 3K3
R_2 = 5K6
R_3 = 12K
R_4;R_5;R_6 = 100K
R_7 = 270K

Capacitors:
C_1 = 220μ; 25 V; axial
C_2 = 10μ; 10 V; radial
C_3 = 1n8
C_4 = 3n9
C_5 = 100n

Semiconductors:
D_1 = 1N4001
D_2 = zener diode 8V2; 400 mW
IC_1 = MC145027 (Motorola)
IC_2 = 4051
IC_3;IC_4 = ULN2001A

Miscellaneous:
PCB Type 87291-1 (see Readers Services page).

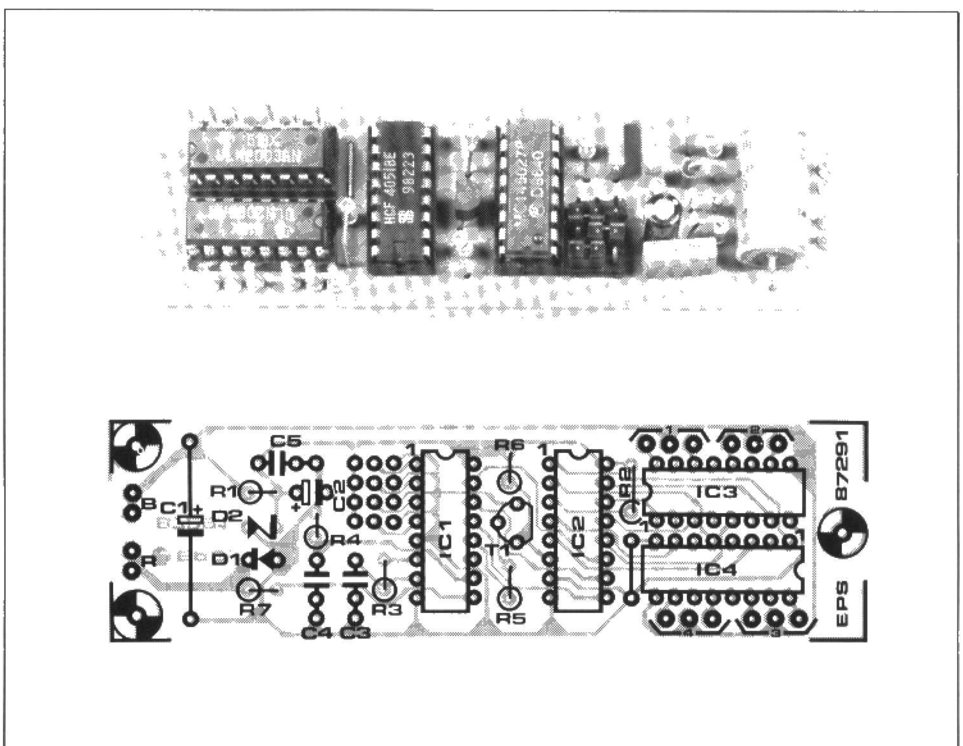


Fig. 6. The printed-circuit board for the decoder.

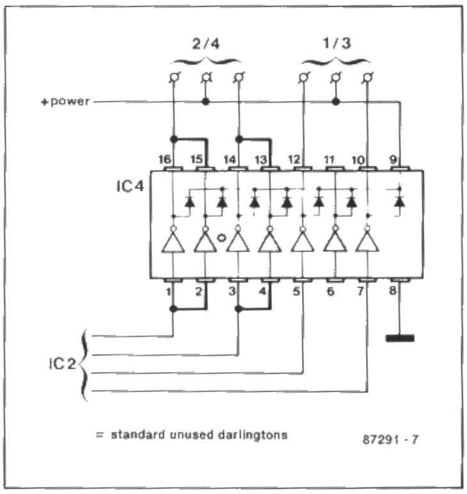


Fig. 7. The maximum current at an output may be doubled by connecting the two relevant darlingtons in parallel with the aid of two short wire bridges.

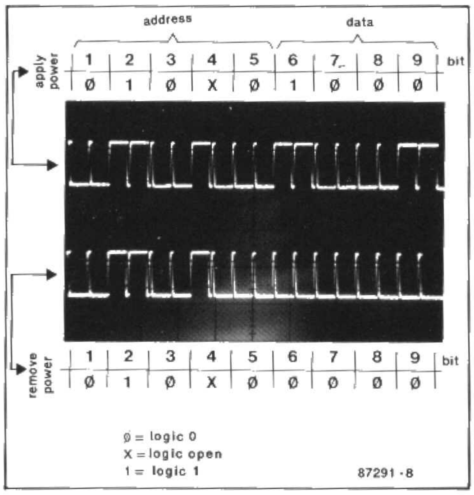


Fig. 8. Actual data streams. The one at the top is generated when bit 9 is set (power applied); the other when the data bits are reset (power removed).

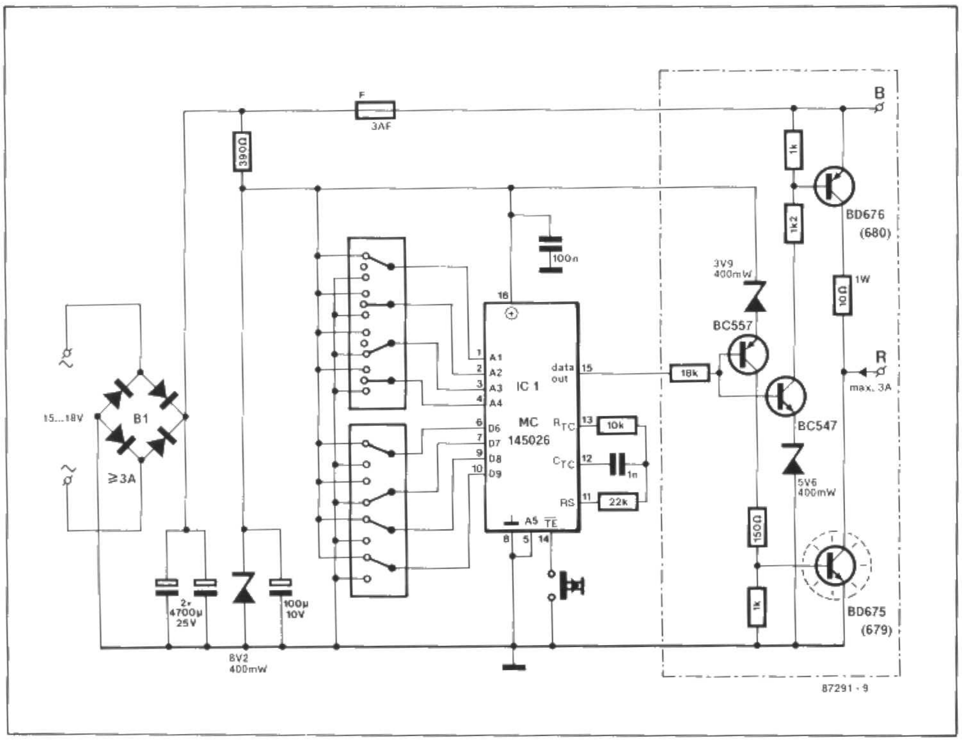


Fig. 9. Circuit diagram of an encoder based on Motorola's MC145026 that enables the Marklin decoder to be used independently.

keyboard number	keyboard DIL switches on	points (turnouts) number	decoder jump wires placed at
1	-	1, 2, 3, 4	2 3 5 7
		5, 6, 7, 8	3 5 7
		9, 10, 11, 12	1 4 5 7 -
		13, 14, 15, 16	2 - 4 5 - 7 -
2	1 - -	1, 2, 3, 4	4 5 7
		5, 6, 7, 8	1 - - - 5 7 -
		9, 10, 11, 12	- 2 5 - 7
		13, 14, 15, 16	- - 5 7 -
3	- 2 - -	1, 2, 3, 4	1 - 3 - 6 7 -
		5, 6, 7, 8	- 2 3 - - 6 7 -
		9, 10, 11, 12	- 3 - 6 7 -
		13, 14, 15, 16	1 - 4 - 6 7
4	1 2 - -	1, 2, 3, 4	2 - 4 6 7 -
		5, 6, 7, 8	- - 4 - 6 7 -
		9, 10, 11, 12	1 - - - 6 7 -
		13, 14, 15, 16	- 2 - - 6 7
5	- 3 -	1, 2, 3, 4	- - - - 6 7 -
		5, 6, 7, 8	1 3 - - - 7 -
		9, 10, 11, 12	2 3 - - 7 -
		13, 14, 15, 16	- - 3 - - - 7
6	1 - 3 -	1, 2, 3, 4	1 - - 4 - 7 -
		5, 6, 7, 8	- 2 - 4 - - 7
		9, 10, 11, 12	- 4 - 7 -
		13, 14, 15, 16	1 - - - - 7 -
7	- 2 3	1, 2, 3, 4	- 2 - - - - 7 -
		5, 6, 7, 8	- - - - - - 7 -
		9, 10, 11, 12	1 - 3 - 5 - 8
		13, 14, 15, 16	2 3 - 5 - - 8
8	1 2 3 -	1, 2, 3, 4	- - 3 - 5 - 8
		5, 6, 7, 8	1 - - 4 5 - - 8
		9, 10, 11, 12	- 2 4 5 - - 8
		13, 14, 15, 16	- - 4 5 - 8
9	- - 4	1, 2, 3, 4	1 - - 5 - - 8
		5, 6, 7, 8	2 - - 5 - 8
		9, 10, 11, 12	- - 5 - - 8
		13, 14, 15, 16	1 - 3 - - 6 - 8
10	1 - - 4	1, 2, 3, 4	- 2 3 - 6 - 8
		5, 6, 7, 8	- 3 - - 6 - 8
		9, 10, 11, 12	1 - - 4 6 - 8
		13, 14, 15, 16	- 2 - 4 - 6 8
11	- 2 - 4	1, 2, 3, 4	- - 4 - 6 - 8
		5, 6, 7, 8	1 - - - 6 - 8
		9, 10, 11, 12	2 - - - 6 - 8
		13, 14, 15, 16	- - - - 6 - 8
12	1 2 - 4	1, 2, 3, 4	1 - 3 - - - - 8
		5, 6, 7, 8	- 2 3 - - - - 8
		9, 10, 11, 12	- 3 - - - 8
		13, 14, 15, 16	1 - - 4 - - - 8
13	- - 3 4	1, 2, 3, 4	- 2 - 4 - - - 8
		5, 6, 7, 8	- - - 4 - - - 8
		9, 10, 11, 12	1 - - - - - 8
		13, 14, 15, 16	- 2 - - - - 8
14	1 - 3 4	1, 2, 3, 4	- - - - - - 8
		5, 6, 7, 8	1 3 - 5 - -
		9, 10, 11, 12	- 2 3 - 5 - - -
		13, 14, 15, 16	- - 3 - 5 - -
15	- 2 3 4	1, 2, 3, 4	1 - - 4 5 - - -
		5, 6, 7, 8	- 2 - 4 5 - - -
		9, 10, 11, 12	- - - 4 5 - - -
		13, 14, 15, 16	1 - 5 - -
16	1 2 3 4	1, 2, 3, 4	- 2 - - 5 - -
		5, 6, 7, 8	- - - 5 - -
		9, 10, 11, 12	1 - 3 6 - -
		13, 14, 15, 16	2 3 - 6 - -

Table 1. Address settings for the Marklin keyboard and the present decoder.

solenoids. The darlingtons are capable of switching up to 500 mA per coil. If points (turnouts) are connected in parallel, this maximum current must be borne in mind. Since not all the darlingtons in IC3 and IC4 are used, it is possible to increase the current from some outputs by a factor 2. To do this, two darlingtons connected to the relevant output (see Fig. 7) are connected in parallel with the aid of two short wire bridges.

If Marklin points (turnouts) with lights are used, these lights are powered by disconnecting the yellow wire from the central terminal of the solenoids and connecting it instead to the central rail. The same may be done with the signal lights. This arrangement will cause the lights to be on permanently. It is worth considering connecting the wire, perhaps via a switch, direct to the yellow AC connection on the transformer. This has the added advantage that the load on the central unit is decreased so that more power becomes available for the trains.

Testing

For testing at least a Marklin central unit, a keyboard, and a "control 80" are needed. When every unit has been connected, a red LED on the central unit will light (it may be necessary to press the GO key on the "control 80" unit first). The connected points (turnouts) may be operated via the keyboard. It is, of course, essential that all addresses are set on the decoder as well as on the DIL switches at the rear of the keyboard (see Table 1).

Every time a key is depressed, two se-

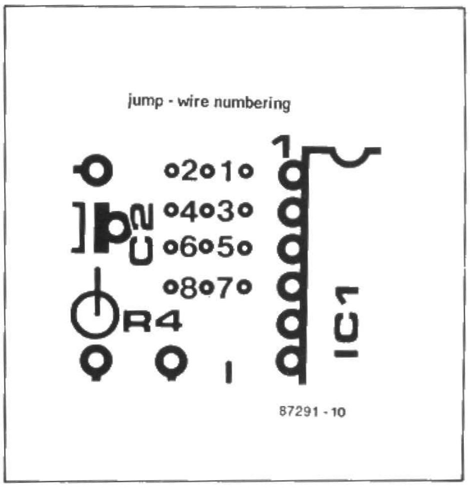
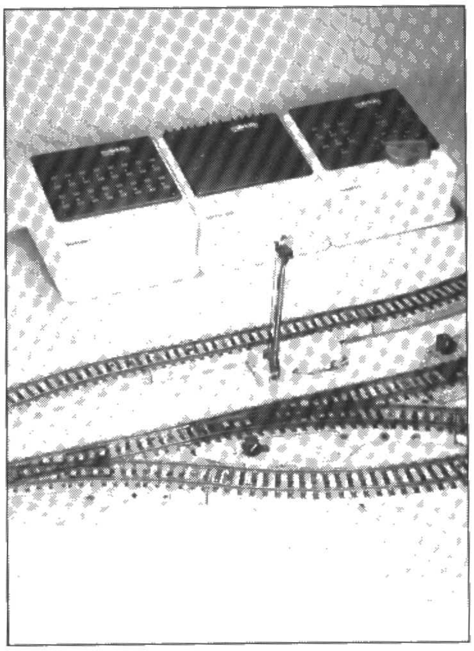


Fig. 10. Jump wire numbering.

quential pulse trains are put on to the rails. Each of these trains contains two identical data bytes. One pulse train would, therefore, be sufficient, because the decoder needs only two data bytes, but for absolute security of operation two trains have been arranged. Bits 1 to 4 constitute the relevant decoder address; bit 5 is always 0; bit 6 to 8 form the sub-address of the appropriate solenoid; and bit 9 is 1 so that power is applied. When the key is released, four bytes are again put on to the rails, but this time with the data bits reset to cause power to be removed.

Alternative control circuit

The points (turnouts) decoder may also be used independent of the Marklin system, not only with model railways, but also as a two-wire remote control unit. This is made possible by Motorola's MC145026 encoder. This IC makes it possible to construct a substitute for the Marklin control with only a few additional components (see Fig. 9). The encoder has 9 address/data inputs. Input



5 is connected to earth. The trinary decoder address must be placed on inputs 1 to 4 and the solenoid sub-address (binary) on inputs 6 to 8. The power is removed or applied by bit 9. In the circuit, switches are shown for setting the addresses, but the inputs may just as well be controlled by the output port of a computer. Note, however, that these inputs are not TTL compatible (not even if the decoder would operate from a 5 V supply). Furthermore, for each address bit a third logic state (high impedance) must be available. This means that an interface is required between the output port and the encoder.

A short pulse at the TE input (transmit enable) results in the set byte being sent twice in succession. If the TE input is kept low permanently, the encoder sends continuously.

When one central power supply is used (as in the Marklin system), the data are superimposed on the supply voltage which is effected by the boxed section at the right in Fig. 9. It is, however, also possible to give each decoder a separate supply, so that only earth and a signal line have to be provided. The boxed section in Fig. 9 is then not required. The output of the encoder and earth are then connected direct to R and the data input respectively (it may be necessary to remove R_7).

EVENTS

Call for papers

Synopses are now requested for the IEE International Conference on Expert Planning Systems which will be held in Brighton from 27 to 29 June 1990.

The conference is intended for all those with an interest in, or responsibility for, planning in business, engineering, finance, or other environments in which complexity is present.

Further information from **Conference Services • IEE • Savoy Place • LONDON WC2R 0BL.**

Synopses of 200—300 word (not for publication) for the 6th BEAMA International Electrical Insulation Conference to be held from 21 to 24 May 1990 in Brighton should be submitted before 31 March this year to **BEAMA • 8 Leicester Street • LONDON WC2H 7BN.**

The **1989 SMARTEX Conference and Exhibition** will be held at the Wembley Exhibition Centre, London, from 7 to 9 February. Further information from the organizers, **MGB Exhibitions Ltd • Marlowe House • 109 Station Street • SIDCUP DA15 7ET • Telephone 01—302 7205.**

The Which Computer? Show is taking place at the National Exhibition Centre, Birmingham from 21 to 24 February. Further information from **Cahners Exhibitions Ltd • Chatsworth House • 59 London Road • TWICKENHAM TW1 3SZ • Telephone 01—891 5051.**

EMC '89, another in ERA's series of seminars on electromagnetic compatibility will be held at the Heathrow Penta Hotel, London, on 7 February. Further information from **ERA Technology Ltd • Cleeve Road • Leatherhead KT22 7SA • Telephone (0372) 374151.**

Asia Telecom 89, the Special Session of the World Telecommunication Forum and Specialized International Telecommunications Exhibitions, will be held in Singapore from 20 to 25 February. Further information from the **ITU, Geneva, or the Telecommunication Authority of Singapore • 31 Exeter Road 2600 • Comcentre Singapore 0923 • Republic of Singapore • Telephone +65 730 3283.**

Saudi Elenex 89, the 2nd Electrical and Electronic Engineering show, will take place at the Riyadh Exhibition Centre from 12 to 16 February. Further information from **Overseas Exhibition Services Ltd • 11 Manchester Square • LONDON W1M 5AB • Telephone 01—486 1951.**

Domotechnica, the domestic appliance and components exhibition, will be held in Cologne, Federal Germany, from 14 to 17 February. Further information from **BEAMA • Leicester House • 8 Leicester Street • LONDON WC2H 7BN • Telephone 01—437 0678.**

A number of seminars on **Data Communications and Telecommunications, and on Information Technology** will be conducted this month by **Frost & Sullivan**. Further information from that organization at **4 Grosvenor Gardens • LONDON SW1W 0DH • Telephone 01—730 3438.**

The Institution of Electrical Engineers (IEE) is to hold its **Third Vacation School on Radiowave Propagation** from 5 to 10 March at the Danbury Management Centre, Chelmsford.

The purpose of the course is to provide an appreciation of the pertinent characteristics of the ionosphere, troposphere, and terrain, and of their effects on practical radio wave communications. The understanding of propagation phenomena is fundamental to the design of communication, radar, remote sensing, and broadcasting systems.

Zoë Bartlett • IEE • Savoy Place • LONDON WC2R 0BL • Telephone 01—240 1871 Ext 308

Elenex Australia '89 will be held in Sydney from 14 to 17 March. Further information from **BEAMA • Leicester House • 8 Leicester Street • LONDON WC2H 7BN • Telephone 01—437 0678.**

Recognizing speech in noise

by Dr William Ainsworth, Department of Communication and Neuroscience, University of Keele

Few people have any experience of communicating verbally with computers and even fewer have ever done so in a noisy environment. Yet in a factory or when using a telephone in a busy office, recognizing and decoding speech is a familiar problem. But it will take many years of research before the most efficient form of man-machine interface will be evolved, though the task has to be tackled if we are to be able to talk to computers against a background of machinery, in a motor car or on a flight deck. Headway is already being made in analysing the difficulties and outlining ways to overcome them.

Speech dominates human communication. If we want people to do something, or we need certain information from them, we simply speak to them. If they are far away we may write them a letter, but most people prefer to pick up a telephone, perhaps because reading and writing seem much more complicated than speaking and listening. That is hardly surprising, for it takes years of practice at school to become proficient in the skills needed to read and write.

When we want to communicate with a machine we have to learn new skills. We need to know how to poke at a keyboard with our fingers and to watch the effect it has on a screen. How much easier it would be if we could simply speak into a microphone to get the machine to do what we wanted!

This dream occurred to speech technologists many years ago, and for the last 20 years or so they have been trying to devise techniques for getting machines to respond effectively to speech signals.

Speech communication appears to be a simple process. It is learned by every healthy child with little or no effort. In reality it is not simple: it is a most complex process. An idea in the mind of the speaker must first be expressed as a sentence in a language understood by both him and the listener. It must then be articulated. We do it by modulating the airstream from the lungs by the vocal cords to produce a sequence of pulses whose frequency determines the intonation. The pulses excite the resonances of the vocal tract and then radiate from the lips as a sound wave. The meaning of the sentence is coded in this wave by subtle movements of the tongue, jaw and lips. These complex movements are known intuitively by everyone who has learned the language.

But this is only half the story. The sound wave passes through the outer ear of the listener and causes the eardrum to vibrate. These vibrations cause the ossicles, a series of small bones attached to the eardrum, to move and pump fluid in the cochlea, or inner ear. In the cochlea is the basilar membrane which oscillates at various places along it which depend upon the frequencies present in the input signal. So, the structure of the inner ear begins the process of decoding the speech wave. Attached to the basilar membrane are a large number of hair cells, some 30,000 of them, which actuate nerve cells when they bend. These cells are the first stage in a complex system which leads up the brainstem and eventually to the auditory cortex.

Automatic recognition

So far, the processes by which the speech signals are decoded by the brain are not well understood, so programming a computer to recognize speech in the same way that the brain operates is obviously impossible. Nevertheless, for many practical purposes a machine which recognizes just a few words can be very useful. For example, consider a program that displays the choices available to the user by means of numbered menus. If the machine can just recognize the spoken digits the user can complete his task by voice.

Most practical speech recognizers work by pattern matching. The user speaks all the words in the machine's vocabulary and the machine analyses them and stores the result. These stored patterns are often known as templates. When an unknown word is spoken, the machine compares this new utterance with each of the stored templates and chooses the one which gives the best match.

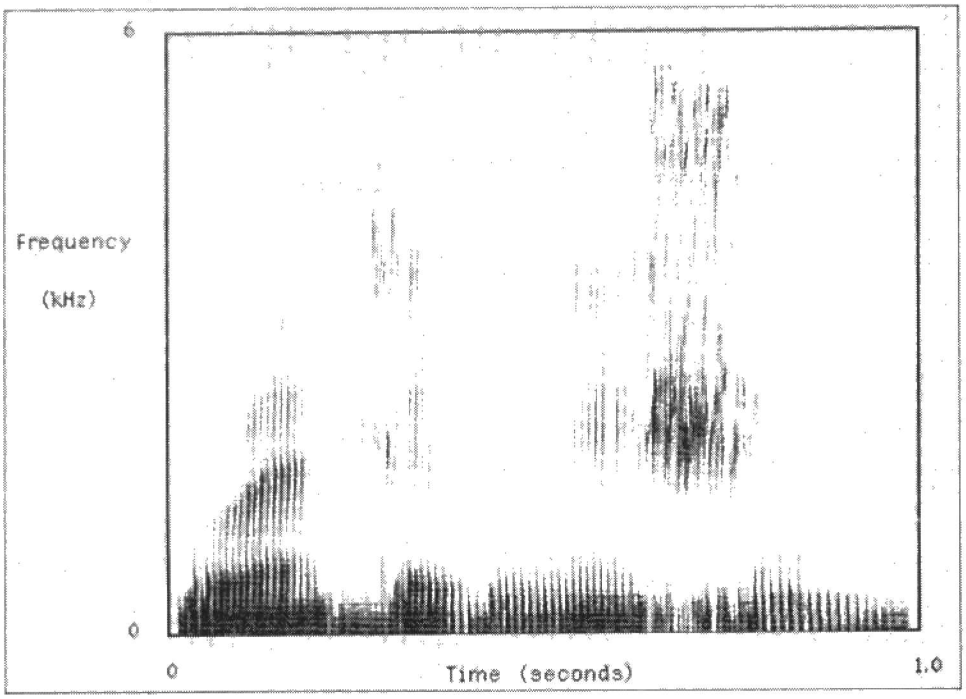
Several techniques have been employed to analyse speech signals. We know that speech is encoded in terms of frequencies and that the human auditory system begins its analysis of sounds by separating them into their component frequencies, so spectral analysis is a popular technique. The upper of the two illustrations shows a sound spectrogram, or sonagram, of the word 'recognition'. Frequency is represented by the blackness of the picture. The dominant frequencies, known as the formants, can be seen as the black horizontal bands. These reflect the resonances of the vocal tract.

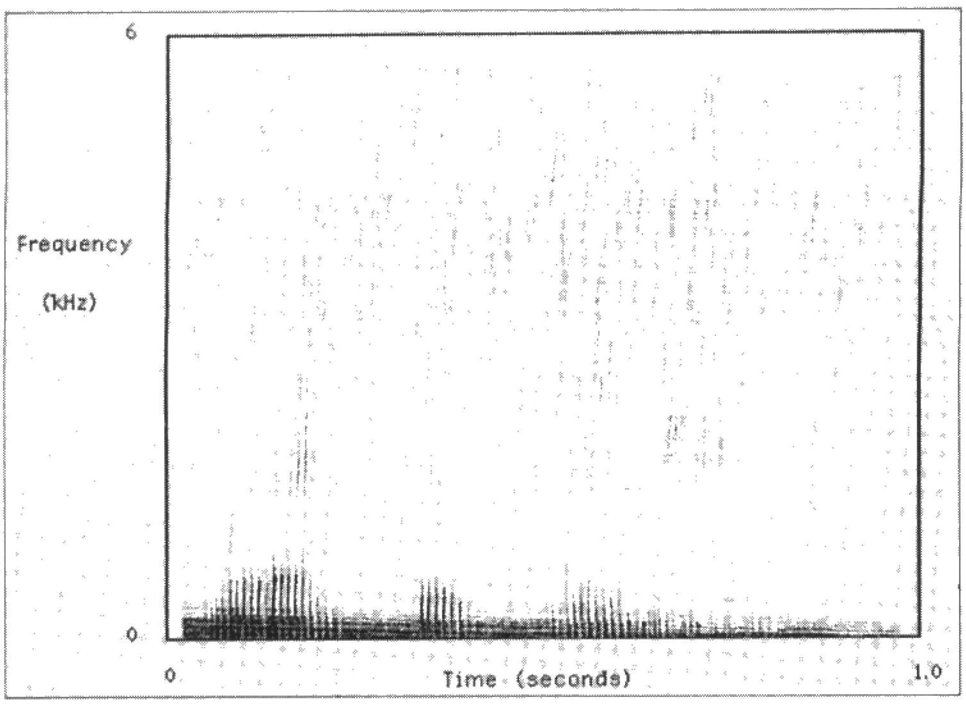
Problems

Speech recognizers built on these principles alone are not very successful for three reasons:

(1) Every time we utter a word we speak at a different rate, so some patterns are spread out in time compared with others.

(2) Different people have different sized vocal tracts, so the formants occur at different frequencies when they say the





Sonagrams of the word 'recognition'. At the top it is spoken against a quiet background, and above with a signal-to-noise ratio of −6 dB.

same word.

(3) Most speech communication takes place not in isolation, but against a background of other noises.

Various techniques have been devised for dealing with these problems. The first problem can be dealt with by so-called dynamic time warping. This enables the stored templates to be expanded or compressed in such a way that the optimum match is obtained. Alternatively the problem can be dealt with by building statistical models of each word which incorporate the variability of the utterances.

Usually the multi-speaker problem has been circumvented by training the system with the voice of the user but there have been some attempts to cope with it by building transformations for each new speaker that enable his voice to be transformed into one like that of the person who originally trained the system. Here, statistical modelling of the variability has again been used.

The problem of recognition in noise has not yet been solved. John Bridle and his colleagues at the UK Royal Signals and Radar Research Establishment in Malvern some years ago showed that a speech recognizer which worked well in the quiet recognized only about 50 per cent of spoken digits correctly when the signal-to-noise ratio was +3 dB (decibels). This is far worse than human performance. It has been known for many years that spoken digits can be recognized with almost complete accuracy with a signal-to-noise ratio as poor as −6 dB, which means the intensity of the speech is much less than that of the noise. A sonagram of the word 'recognition' with a signal-to-noise ratio of −6 dB is shown in the lower of the two illustrations.

Auditory modelling

The superior performance of people in recognizing speech in noise has led to the suggestion that speech analysers which operate on the same principles as the human auditory system might work better than those based on conventional techniques. Preliminary experiments by Dr Ghitza at the Bell Laboratories in the USA and others elsewhere have shown promising results.

Our Department of Communication and Neuroscience comprises a number of research groups which investigate the mechanisms of vision, hearing and speech. Professor Ted Evans, the head of the department and leader of the Auditory Physiology group, has developed an electronic model of a single channel of the auditory system. It gives responses similar to those obtained by inserting micro-electrodes in the auditory systems of cats.

Professor Evans' model consists of a filter with characteristics that simulate those of the inner ear, a half-wave rectifier and logarithmic compressor to represent the action of the hair cells, and what is called a probalistic spike generator to simulate the production of action potentials in nerve cells.

Our Speech and Auditory Physiology groups are collaborating with Dr Pat Wilson of the Auditory Psychophysics group to produce a computational model of the auditory system with 100 or more channels. This work is made possible by a grant from the UK Science and Engineering Research Council to install a fast computer that will enable the model to process signals, especially speech, in a reasonable time.

The first stage of the model consists of a bank of bank-pass filters which simulate the signal processing as far as the auditory nerve. The characteristics of these filters are estimated by a process known as reverse correlation. A random noise signal is applied to the auditory system and responses are recorded from the auditory nerve by means of a microelectrode. The noise signal causing the nerve fibre to respond is also recorded. By a process similar to cross correlation between the noise input signal and the response of the nerve fibre, the impulse response (the response of a filter to a single impulse) of the auditory filter is found (in practice the impulse response is reversed in time; hence the term reverse correlation). Several ex-

periments have to be done with a number of nerve cells, each tuned to respond to different frequencies, to develop the impulse responses of a bank of filters.
These impulse responses can be programmed on the computer and used to simulate the filtering characteristics of the auditory system. The other stages of auditory processing, logarithmic compression and rectification by the hair cells and the generation of spikes according to a probability function, can also be programmed. The result is a computational model which allows the signals generated at each level in response to speech sounds to be studied.
The auditory system is more complicated than I have already outlined. Recent physiological studies have shown that there are interactions between the channels: if there is activity in one channel, the activity in neighbouring channels is suppressed. This mechanism might be responsible for reducing the effects of noise while enabling speech signals to be transmitted to the higher regions of the auditory system. We intend to build lateral suppression into our model and to investigate what effect it has on speech processing.

Speech synthesis

Techniques for speech synthesis were developed about 20 years ago. In a typical system a sentence is first translated into a sequence of phonetic units, which represent the way in which each sound is pronounced. This can be done by looking up each word in a phonetic dictionary or by applying a set of context-sensitive rules (for example *p* followed by *h* is pronounced *f*, otherwise *p*).
The phonetic units are then translated into acoustic parameters which represent the physical characteristics of the sounds. The acoustic parameters are the frequencies of the formants (see left-hand illustration), their intensities, and their durations. They are used to control a speech synthesiser consisting of a set of resonators excited by a sequence of pulses.
Although such a system produces intelligible speech, the output sounds rather mechanical. Moreover, it has been found that when it is heard against a background of noise it is a great deal less intelligible than equally loud natural speech. We are collaborating with the IBM Scientific Centre in Winchester to try to discover why this is so.
One possibility is that whereas this system faithfully models the resonances of the vocal tract it does not employ realistic excitation pulses. A technique known as inverse filtering is being used to measure the shapes of the excitation pulses in human speech. In this technique the characteristics of the vocal tract filter are estimated, and then the characteristic of the filter is inverted. If speech signals are passed through the 'inverted' filter, only the excitation pulses remain.
Using the technique we are able to study the variation in shape of the excitation pulses. This knowledge can be applied to speech synthesis. We expect that speech synthesized in this way will be more intelligible in the presence of background noise.

User interface

When the captain of a ship gives a compass course for the helmsman to steer, the helmsman repeats it back to confirm that he has heard it correctly. When a telephone operator is asked to obtain a number she repeats the number back. Communicating with a machine in a noisy environment is somewhat similar. The noise may corrupt the speech signal and cause an error in recognition. The user will be unaware of the mistake unless the words are displayed on a screen or the machine is equipped with a synthesizer to speak back to him. If the user is communicating over a telephone line, or if his eyes are busy whith another task, the latter course may be the only one that is practicable.
The question arises as to whether the response of the recognizer should be checked after each word has been spoken to it or whether it should be checked later, for example, at the end of each sentence. Compass courses always consist of three digits and they are repeated back as a group. Telephone numbers, on the other hand, vary quite widely in the number of digits they contain. They are often checked after three digits, but on a bad line digits may be checked one by one.
Here at Keele we are interested in communicating with computers in a noisy environment where it is likely, in spite of advances in recognition from auditory modelling and in synthesis from realistic excitation pulses, that occasional mistakes will be made. So we are interested in finding the most efficient ways of detecting and correcting errors. We have developed a mathematical model of the user interface, which enables us to arrive at the optimum number of words which should be spoken before any checking is done. This model predicts, as might be expected, that as the noise level rises and the frequency of errors increases, the number of words spoken before a check is made should be reduced. Experiments have shown that the specific predictions of the model are borne out in practice.

Future plans

Our research is by no means completed. We have only recently acquired computers powerful enough for us to carry out the work. When even more powerful computers come into use, progress will be faster.
Advances are continually being made in understanding the physiology of the auditory system. We intend to incorporate these developments in future auditory models and to test their utility in automatic speech recognizers. Our programme in speech synthesis has been hampered by a lack of knowledge as to how the shape of the excitation pulses varies in natural speech. We are gradually acquiring this knowledge and in due course it will be transferred to our speech synthesis system.

NEWS

Mainland Europe base for Marconi Instruments

Marconi Instruments is to establish its first major research, development, and manufacturing base in mainland Europe as a result of buying the French electronics company Adret.
It will be run by a new company, called Marconi—Adret SA, which will be MI's first overseas subsidiary. MI already has seven overseas sales and service offices, and more than 130 international representatives.

Award for 'MAC' development team

The development team at the IBA, which created the MAC television system to be used by all European direct-to-home broadcast satellite systems, has been awarded the 1988 coveted JJ Thomson medal for its outstanding contribution to electronics.
The JJ Thomson Medal is awarded annually by the Institution of Electrical Engineers (IEE) for an outstanding contribution by a person or a group of persons in electronics, theory, development, or manufacture. JJ Thomson (1856-1940) was awarded the Nobel Prize for Physics in 1906. He worked on the theory of X-rays and atomic theory and was involved with the discovery of the electron.

European Intel repair centre

Intel has underlined its commitment to worldwide customer support with the opening of a new European, purpose-built, microsystem repair centre at its European headquarters in Swindon.
The half-million dollar centre, which greatly increases the company's repair facilities, is capable of supporting all Intel architecture and is their largest 386TM repair centre in Europe for both systems and board products.

NEW BOOKS

CORPORATE INFORMATION MANAGEMENT

by Ivan F. Jackson
ISBN 0 13 174236 1
338 pages — 235×177 mm
Price £15.95 (soft cover)

This book was prompted by the considerable changes that have occurred in the corporate computer environment during the past few years. It is now recognized that the way computers work is of negligible interest: instead, the prime concern is to deliver the right information to the right place at the right time.

Although the book is about corporate information management, it tends, particularly in the first five chapters, to concentrate on the information resource management (IRM) function as the main vehicle for the management of the associated environment, technology and people.

Chapters 6 through 9, dealing with general policy considerations, will be of direct value to the information services manager, especially those sections devoted to financial and human resource management.

The final two chapters are concerned with assessing the performance of the IRM and information services functions, as well as of their managements. The emphasis on performance of an organization as a whole is seen as the key to the performance of the individual activities.

The book is one of the unabridged paperback reprints of established titles widely used by universities throughout the world.

These lower-priced editions are published for the benefit of students by

Prentice Hall • 66 Wood Lane End • Hemel Hempstead HPr 4RG.

ELECTRONIC DEVICES AND COMPONENTS

by J. Seymour
ISBN 0 582 01493 X
575 pages — 233×156 mm
Price £14.95 (soft cover)

The second edition of this comprehensive introduction to the physical principles behind the operation of electronic devices and components has been fully revised to keep pace with modern technology and courses in higher education. Covering the whole range of semiconductor devices, from the p-n junction, through bipolar and field-effect transistors to microwave and power transistors and memory elements, this new edition takes account of important developments in the field and includes:

- computer simulation of semiconductor devices using specially written SPICE programs, complete with program listings;
- expanded and updated information on optoelectronics and optical fibre communications;
- details of gallium-arsenide ICs;
- numerous worked examples and 'points to remember' in each chapter.

The book provides a complete course in electrical and electronic engineering or applied physics for those studying for CNAA degrees. It is also suitable for students preparing for the examinations of the Council of Engineering Institutions and the Institute of Physics. Moreover, it may serve as a self-study aid for practising engineers and technicians.

Longman Scientific & Technical • Longman House • Burnt Mill • HARLOW CM20 2JE

DESIGNING ASICS

by Paul Naish and Peter Bishop
ISBN 0 7458 0626 0
188 pages — 240×165 mm
Price £12.95 (soft cover)

Designing ASICs is a book in the Ellis Horwood series in Electrical and Electronic Engineering and is aimed at practising electronic engineers who are changing from discrete logic to ASICs, students of electronics and computing science, and researchers into computer architecture, telecommunications and control systems.

Application-specific integrated circuits from the fastest growing sector of the semiconductor market, but there is a dearth of books on this important topic. *Designing ASICs*, therefore, meets an urgent need for a work that sets out the theory and practice of the new design techniques.

The book starts with the fundamentals of CMOS technology and the transition in design principles from the use of discrete logic elements to ASICs based on CMOS. It then sets out the principles of synchronous circuit design and circuit definition which underlie all the design techniques that follow.

The middle portion of the book discusses a sequence of particular circuit elements (including latches, RAM interfaces, communicating asynchronous processes, first-in-first-out (FIFO) buffers, pipelining) and provides useful examples of the implementation of these elements.

The final portion of the book deals with testing and discussing a range of design for test techniques. A set of register transfer models is given to enable the behaviour of circuits to be studied further.

The book includes a very useful glossary of terms and a detailed index.

This is a work that no practising engineer or student can afford to do without.

John Wiley & Sons Ltd • Baffins Lane • Chichester PO19 1UD.

TELEVISION ENGINEERING
Broadcast, Cable and Satellite

Edited by R.S. Roberts
ISBN 0 7273 2104 8 - Part 1
ISBN 0 7273 2105 6 - Part 2
259+260 pages — 240×160 mm
Price £00.00 - Part 1
Price £00.00 - Part 2

Both volumes of *Television and Engineering* are based on the Royal Television Society's Television Engineering Course which the Society has run each year for the past four years. The chapters are based on lectures presented by the authors, many of whom are acknowledged experts in their own field and of international standing.

The contents range from basic fundamentals in Part 1 to specific engineering applications in Part 2. Students of television will need both parts and although practising engineers will be more interested in Part 2, they will, none the less, find Part 1 a valuable reference for fundamentals.

The subject treatment is unique in many cases. For instance, Chapter 2 on 'Modulation' does not confine itself to the types of modulation to be found in any current television system. It deals with modulation in a comprehensive manner that will be basic to any further developments that may take place in the future.

Chapter 8 on 'The NTSC and PAL Systems' is also a comprehensive treatment which is basic to any system that uses a sub-carrier. The NTSC system is dealt with in depth because it is the basis of the later variants such as PAL and SECAM.

Another chapter with special features is Chapter 11 on 'Camera Tubes and the Camera', where the author has brought together a large amount of widely scattered information.

Finally, Chapter 16 on 'The Receiver' is probably the first concise treatment of the modern television receiver.

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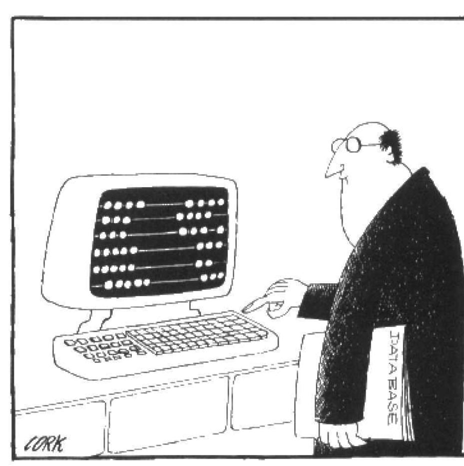
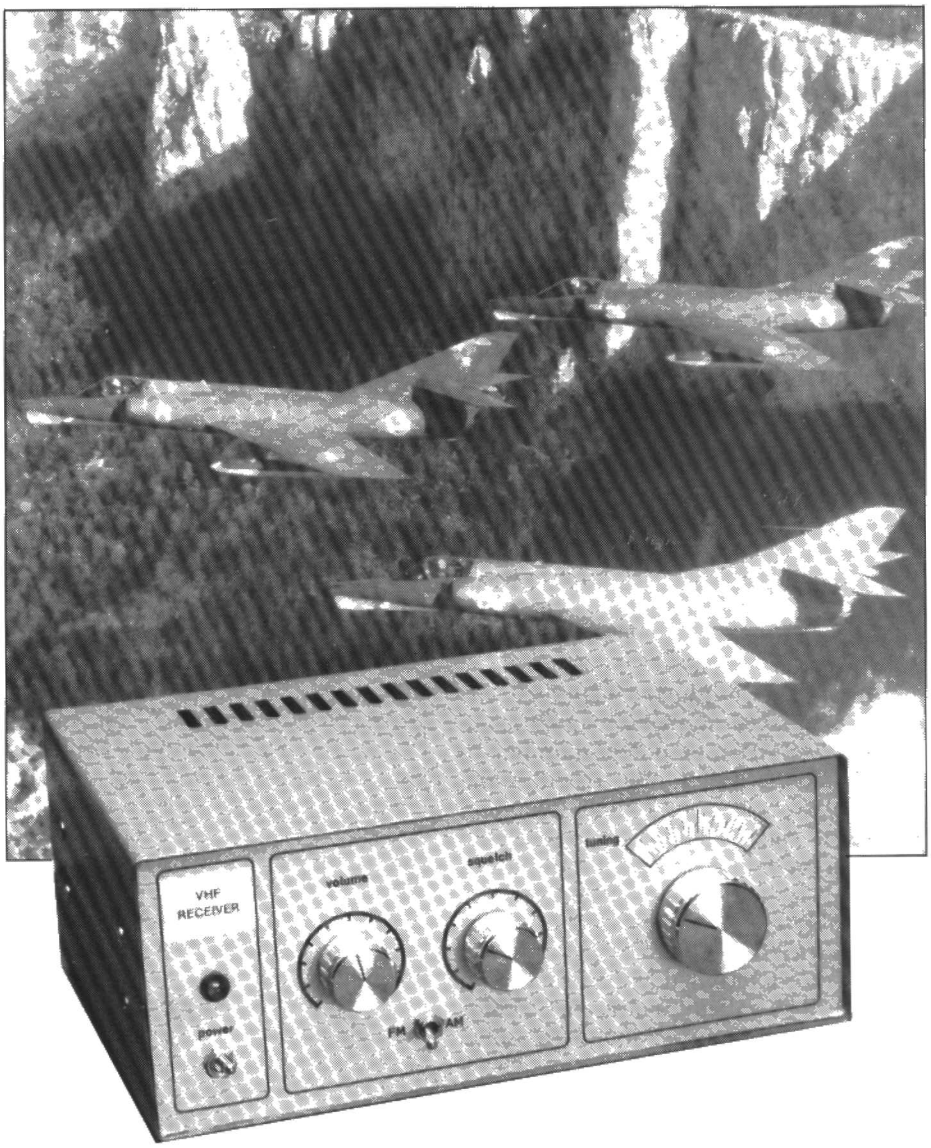


AM/FM VHF RECEIVER

This compact, sensitive, communications receiver has a tuning range of 80 to 135 MHz, covering part of the VHF-low band, the entire VHF broadcast band, and the VHF air band. The upper frequency limit of the receiver can be changed fairly easily to include the 2-metre amateur band.

by J. Bareford



The block diagram given in Fig. 1 shows that the present receiver is a single-conversion type with an intermediate frequency of 10.7 MHz. The RF section, comprising the RF input amplifier, the mixer and the local oscillator, is of conventional structure, and requires no further detailing here. An integrated circuit, to be discussed later, provides the necessary IF amplification, and at the same time comprises the AM and FM demodulators. A squelch (noise suppression) circuit, built from discrete components, works in AM as well as FM mode.

RF circuit

With reference to the circuit diagram of the RF section of the VHF receiver, given in Fig. 2, the aerial signal is raised

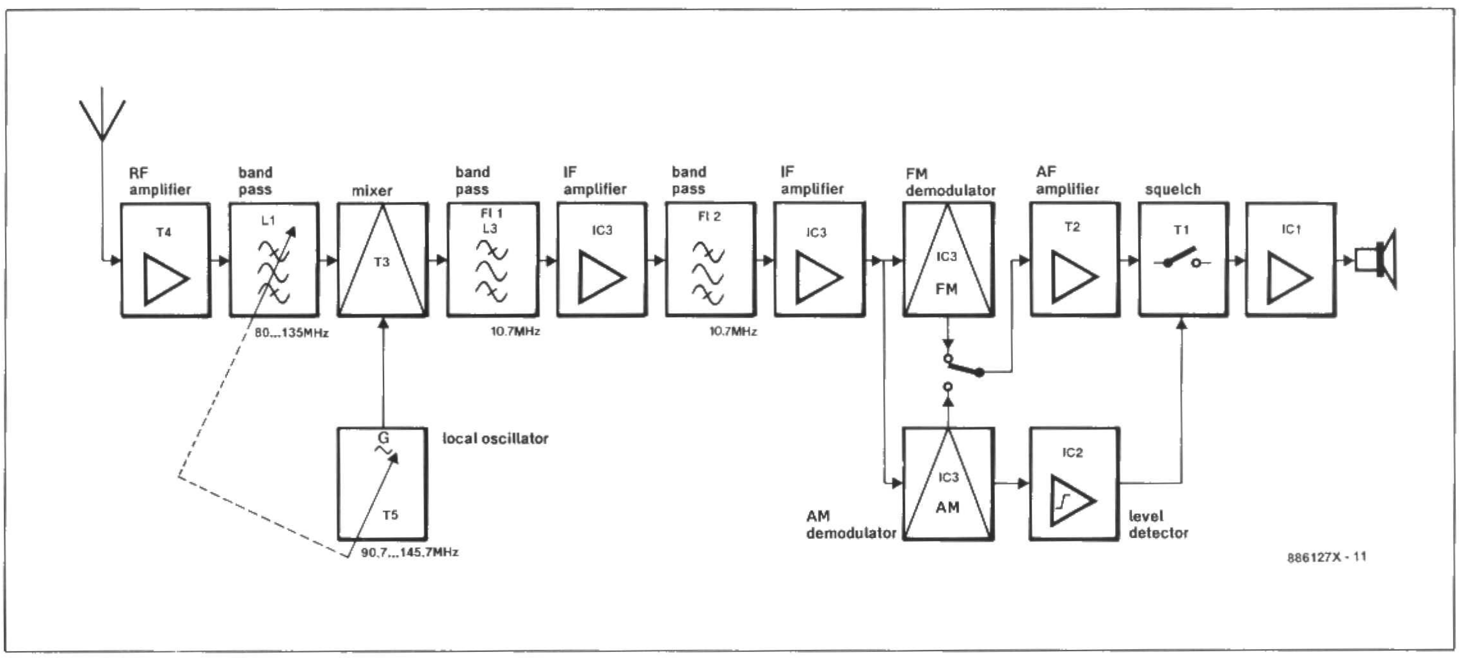


Fig. 1. Block diagram of the AM/FM VHF receiver.

in a wide-band input amplifier based around low-noise transistor Type BFG65. A 10.7 MHz high-pass filter, L_6-C_{20}, is fitted at the aerial input to prevent IF breakthrough. The amplified RF signal available at the collector of T_4 is coupled out to a tap on tuneable band-pass filter L_1-(C_{15}+C_{16}). A double-section tuning capacitor, C_{16}, tunes the band-pass filter together with L-C combination L_2-(C_{16}+C_{24}) in the local oscillator set up around dual-gate MOSFET T_5. The amplified RF signal is applied to gate-1, the LO signal to gate-2, of mixer T_3. The difference frequency, 10.7 MHz, is filtered out with the aid of tuned circuit L_3 in the drain line of the DG MOSFET.

IF circuit and demodulators

Details of the IF amplifier, demodulators, squelch and AF amplifier are given in the circuit diagram of Fig. 3. The signal at point A in the previously discussed RF section is applied to ceramic filter FL_1, which adds to the function of L_3 by reducing the overall bandwidth of the receiver.

Since the Type NE604N integrated circuit combines a number of functions in the receiver, yet may not be familiar to many readers, its internal structure and pinning are given in Fig. 4. The prefiltered IF signal applied to pin 16 is raised in an on-chip amplifier. From output pin 14, it is applied to a second ceramic filter, FL_2, and fed back to the second IF amplifier in the chip. This amplifier drives the internal FM demodulator, which is a so-called quadrature detector that works in conjunction with tuned circuit L_4. Since the on-chip mute-circuit is disabled, the demodulated AF signal is available at pin 6.

The output from the signal-strength detector internal to the NE604N is available at pin 5. Since RF power, and with it RF and IF signal strength, is a function of the amplitude of the modulation signal applied to an AM transmitter, pin 5 of IC_3 simply carries the demodulated AM signal when the receiver is tuned to an AM station (AM is the standard in VHF air-traffic communication).

Squelch and AF amplifier

The FM or AM input signal for buffer T_2 is selected with mode switch S_1. The FET ensures a low driving impedance for the LM386-based AF amplifier, and at the same time prevents the outputs of the NE604N being damaged by the virtual short-circuit to ground when the squelch transistor, T_1, conducts. T_1 is driven by opamp IC_2, which is configured as a comparator, comparing the direct voltage at pin 5 of the NE604N to a level set with potentiometer P_2. When the signal strength of the received trans-

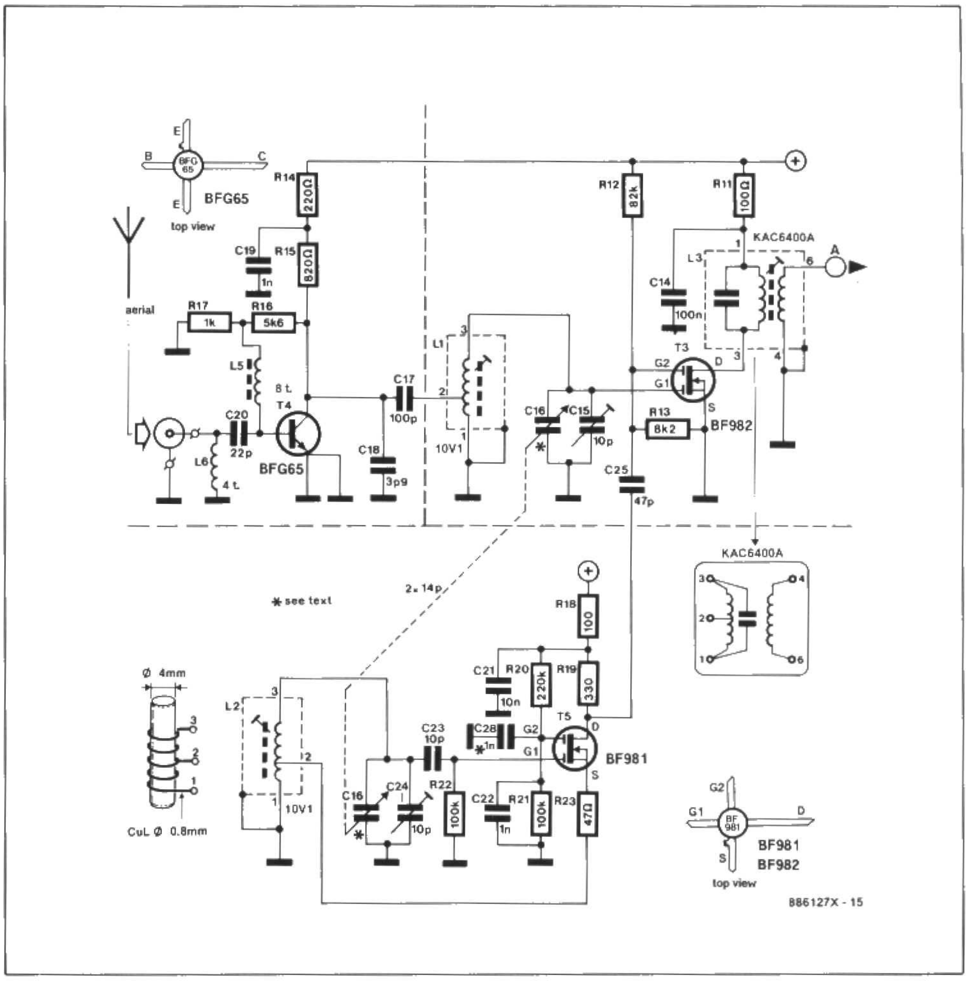


Fig. 2. Circuit diagram of the RF section of the VHF receiver. Note the use of a ganged tuning capacitor, C_{16}.

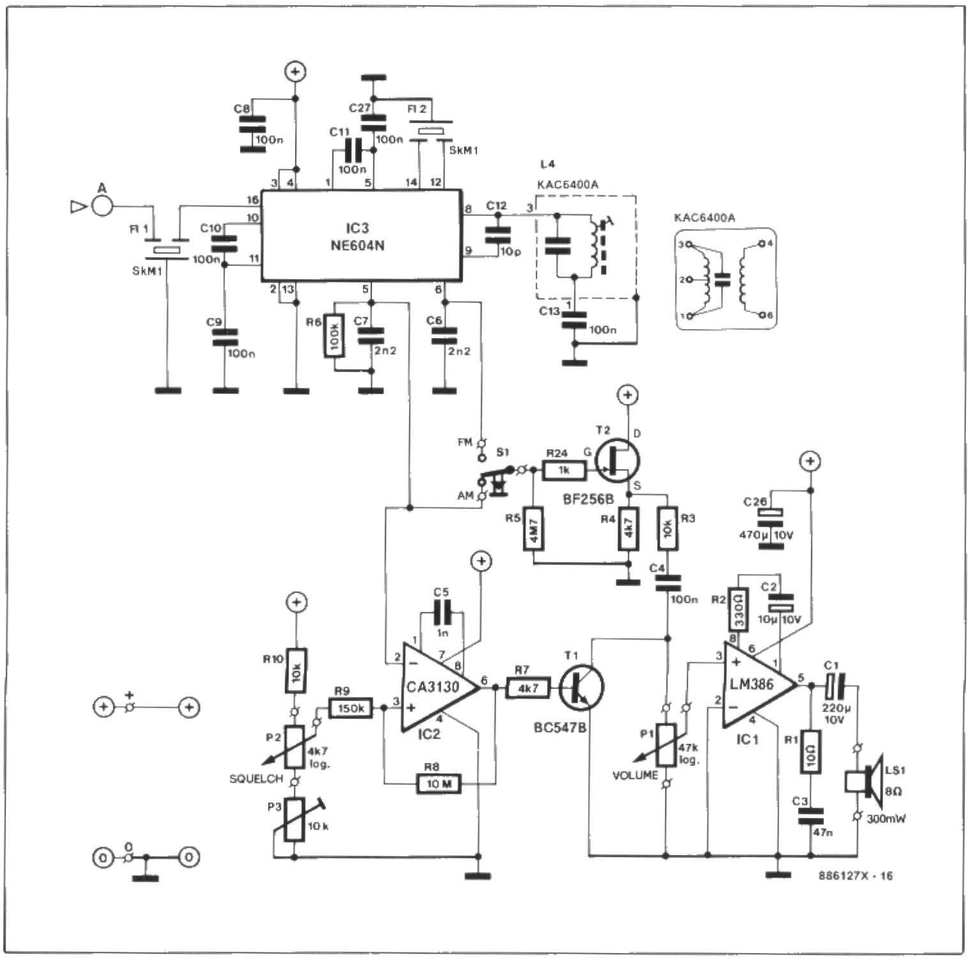


Fig. 3. Circuit diagram of the 10.7 MHz IF section, squelch and AF amplifier.

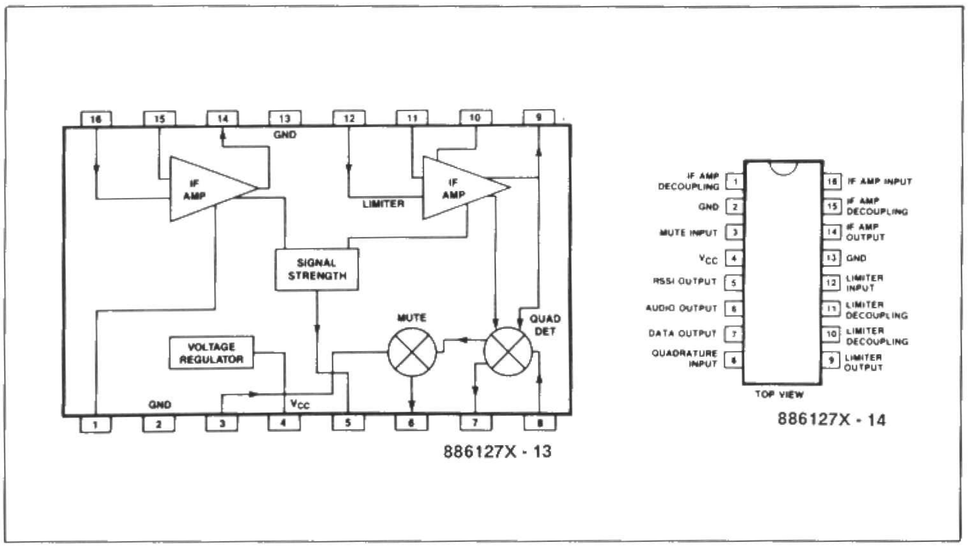


Fig. 4. Internal structure and pinning of the NE604N IF amplifier/FM demodulator from Philips Components.

mitter exceeds the squelch threshold set with P_2, the output of the comparator is virtually 0 V because the voltage at its − input is higher than that at its + input. T_1 is then switched off, so that the demodulated signal is no longer shorted to ground, and can reach AF amplifier IC_1. Resistors R_8 and R_9 provide some hysteresis in the comparator circuit to prevent this toggling as a result of small signal-strength fluctuations. Hysteresis is, of course, essential for AM reception, since without it oscillation would occur. Sometimes, oscillation may still occur, however, and in these cases, the resulting hum level may be reduced by setting P_2 to a slightly higher trip level. Should the comparator persist in oscillation, either increase the value of C_5, or reduce that of R_8.

The AF amplifier set up around the Type LM386 integrated circuit is a standard application, and requires no further detailing.

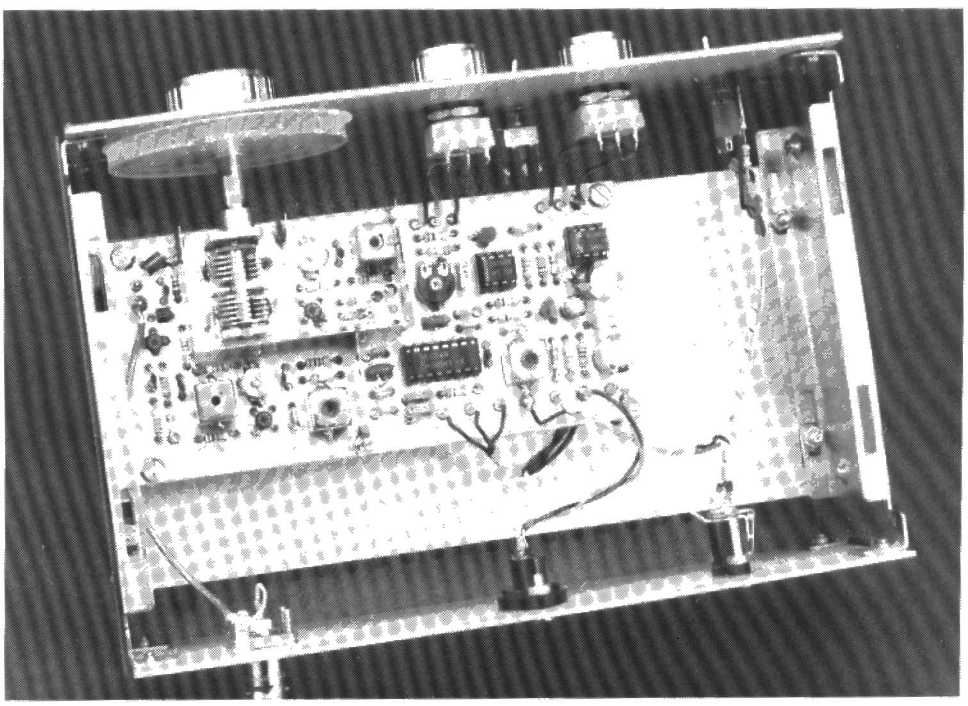
Building the receiver

Start the construction of the receiver by winding inductors L_1 and L_2. These are identical, and wound as shown in the circuit diagram on the white, ABS, former supplied with the Neosid Type 10V1 inductor assembly. Point 2 is a tap made at about 2 turns from the earthy end of the inductor. Great care should be taken not to overheat the base of the former as the wires are soldered to the three pins at one side of the base. Also make sure that the solder joint made on the three connected pins can not cause a short-circuit with the inside of the metal screening can to be fitted later. Check the completed inductors for correct continuity.
Proceed with making L_5 and L_6. The first consists of 8 close-wound turns of 0.2 mm dia. enamelled copper wire; the internal diameter is about 3 mm and no former is used. L_6 is a VHF choke consisting of 4 turns of 0.2 mm dia. enamelled copper wire, wound through a 3 mm long ferrite bead. Carefully remove the enamel coating at the wire ends of these inductors.

The printed-circuit board for the VHF receiver is a double-sided, but not through-plated, pretinned type. The component mounting plan is given in Fig. 5. The component side is left largely unetched to enable it to function as an earth plane. To effect through-contacting and short connections to earth, some component terminals are soldered at *both* PCB sides, i.e. direct to the earth plane at the component side, and to a solder island at the track side of the board. The two rotor connections of PTFE foil trimmers C_{24} and C_{15} are among the component terminals that require soldering at both PCB sides. Every possible care should be taken to solder these terminals as quickly as possible to the ground plane to prevent damaging the foil by overheating.

The rotor plates of the tuning capacitor, C_{16}, are internally connected to the base plate. This is soldered direct to the earth plane, opposite the stator terminals. Short wires are used to connect these to the relevant holes in the PCB. One connection goes to junction C_{23}-C_{24}, the other to junction C_{15}-3(L_1)-G2(T_3).
Surface-mount capacitor C_{28} is not shown on the component overlay of the board. This part, which is essential to ensure oscillator stability, is soldered direct between the G2 terminal of T_5 and the earth plane. Stray radiation from the oscillator and the RF preamplifier is prevented by soldering 20 mm high tinplate or brass screens on to the earth plane, as indicated by the dashed lines on the component overlay.
The IF amplifier/demodulator, IC_3, is soldered direct on to the board, i.e., without an IC socket. Solder pins 2 and 13 to ground as outlined above.

The input to the receiver should be made in thin coaxial wire connected between the two soldering pins and a BNC or SO-239 socket mounted on to the rear panel of the enclosure. The connections to the front-panel mounted squelch and volume potentiometers may be made in unshielded wires, but only if these are kept shorter than about 5 cm.
As shown in the photograph of Fig. 6, the connections to the external loudspeaker and the power supply may be made in a DIN loudspeaker socket and a small DC-input socket, respectively.
A standard 8 V supply, set up around a 7808 with decoupling capacitors at the input and output, may be built into the receiver enclosure. This has the advan-



A look inside the completed prototype.

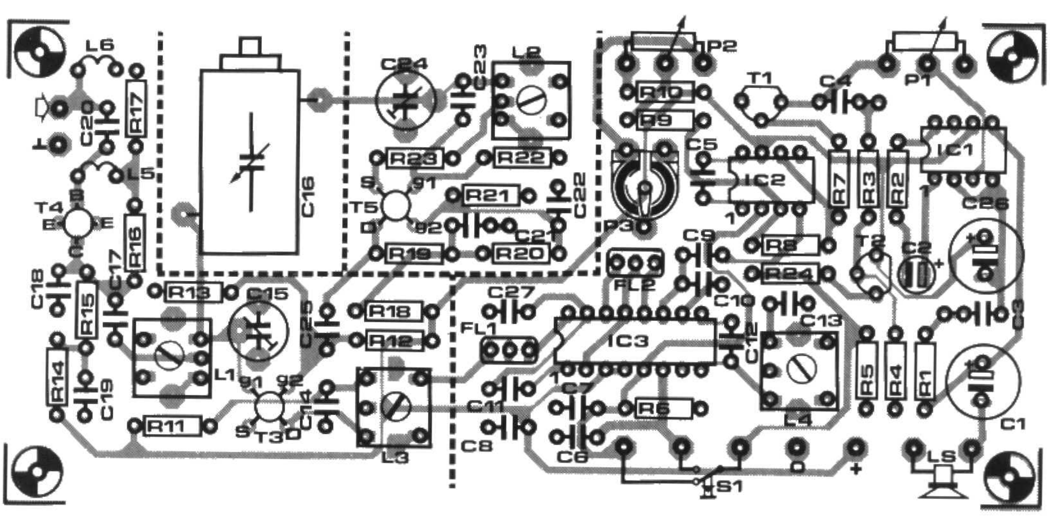


Fig. 5. Printed-circuit board for the VHF receiver.

Parts list

Resistors (±5%):
R_1 = 10R
R_2;R_{19} = 330R
R_3;R_{10} = 10K
R_4;R_7 = 4K7
R_5 = 4M7
R_6;R_{21};R_{22} = 100K
R_8 = 10M
R_9 = 150K
R_{11};R_{18} = 100R
R_{12} = 82K
R_{13} = 8K2
R_{14} = 220R
R_{15} = 820R
R_{16} = 5K6
R_{17};R_{24} = 1K0
R_{20} = 220K
R_{23} = 47R
P_1 = 47K logarithmic potentiometer
P_2 = 5K or 4K7 linear potentiometer
P_3 = 10K preset H

Capacitors:
C_1 = 220µ; 10 V; radial
C_2 = 10µ; 10 V; radial
C_3 = 47n
C_4;C_8;C_{27} = 100n
C_5;C_{19} = 1n0
C_6;C_7 = 2n2
C_9;C_{10};C_{11};C_{13};C_{14} = 100n; ceramic
C_{12};C_{23} = 10p
C_{15};C_{24} = 10p foil trimmer (yellow)
C_{16} = 2×14p ganged tuning capacitor with gearing. Available from Meek-it Elektronika
C_{17} = 100p
C_{18} = 3p9
C_{20} = 22p
C_{21} = 10n ceramic
C_{22} = 1n0 ceramic
C_{25} = 47p
C_{26} = 470µ; 10 V; radial
C_{28} = 1n0 surface-mount capacitor

Semiconductors:
IC_1 = LM386N
IC_2 = CA3130E
IC_3 = NE604N (Philips Components)
T_1 = BC547B
T_2 = BF256B
T_3 = BF982 (C-I Electronics)
T_4 = BFG65 (Universal Semiconductor Devices; C-I Electronics)
T_5 = BF981

Inductors:
L_1;L_2 = Neosid assembly Type 10V1. Winding details are given in the text. (Neosid Limited • Icknield Way West • LETCHWORTH SG6 4AS. Telephone: (0462) 481000. Telex: 826405. Contact: Mr. E. Adcott. Neosid inductors are also available from C-I Electonics, P.O Box 22089 • 6360 AB Nuth • Holland).
L_3;L_4 = KAC6400A (Toko; UK distributor is Cirkit PLC; telephone 0992 441306)
L_5 = see text.
L_6 = see text.

Miscellaneous:
FL_1;FL_2 = SKM1 or similar 10.7 MHz; 50 kHz ceramic filter.
S_1 = miniature toggle switch.
BNC or SO-239 (Amphenol) socket.
Loudspeaker; 8 Ω; min. 300 mW.
PCB Type 886127 (see Readers Services page).
Metal enclosure: approx. size 20×14×8 cm.

tage of allowing the use of an inexpensive mains adapter with 12-18VDC output.

The tuning vernier, of which the design is given in Fig. 7, is glued onto a 5 mm thick perspex disc, drilled in the centre for securing on to the spindle of the tuning capacitor.

It is imperative that the VHF receiver be mounted in a metal enclosure, for which a front-panel is made as suggested in Fig. 8.

Setting up

To begin with, the cores of the four inductor sets are screwed in halfway the formers with the aid of a nylon trimming tool. The two trimmer capacitors, C_{24} and C_{15}, and the squelch range preset, P_3, are set to the centre of their travel. Short-circuit the receiver input. Connect a loudspeaker (min. 8 Ω), and apply power.

First, check the presence of the supply voltage, 8 V, at a number of points on the board. Connect an AC-coupled frequency meter to gate-2 of T_3. Set the tuning capacitor to full capacitance, and adjust L_2 for 90.7 MHz. Adjust C_{24} if this frequency can not be obtained even with the core of L_2 fully screwed in. Set the tuning capacitor to minimum capacitance, and check that the oscillator frequency is about 145 MHz. Now re-adjust C_{24} and, if necessary, the core in L_2, until the desired tuning range is obtained.

Set S_1 to FM mode, and disable the squelch by turning the control fully anticlockwise. Peak L_4, L_3, and L_1 for highest, stable, AF noise output.

Remove the short-circuit at the receiver input, and connect a 50 to 75 Ω unbalanced aerial, e.g., a whip or a ground-plane type. Tune to a relatively strong FM transmission, and adjust the quadrature coil, L_4, until the

demodulated sound is undistorted. Tune to a relatively weak transmission, or attenuate the aerial signal, and re-adjust L_1 and L_3 for minimum noise. This adjustment may also be carried out by switching to AM mode and tuning to an air-band beacon (in many areas, these may be found between 110 and 120 MHz).
Finally, the span of the squelch control may be set to individual requirements by adjusting P_3.

Radio amateurs and other experienced RF constructors will have little difficulty modifying the receiver for a higher maximum frequency, so that the 2-m (144-146 MHz or 144-148 MHz) and weather satellite bands (135-137 MHz) are covered at the expense of shifting the lower tuning limit from 80 to about 90 MHz. Obviously, this requires less inductance for L_1 and L_2, so that some experimentation may be needed in reducing the number of turns and moving the taps accordingly.

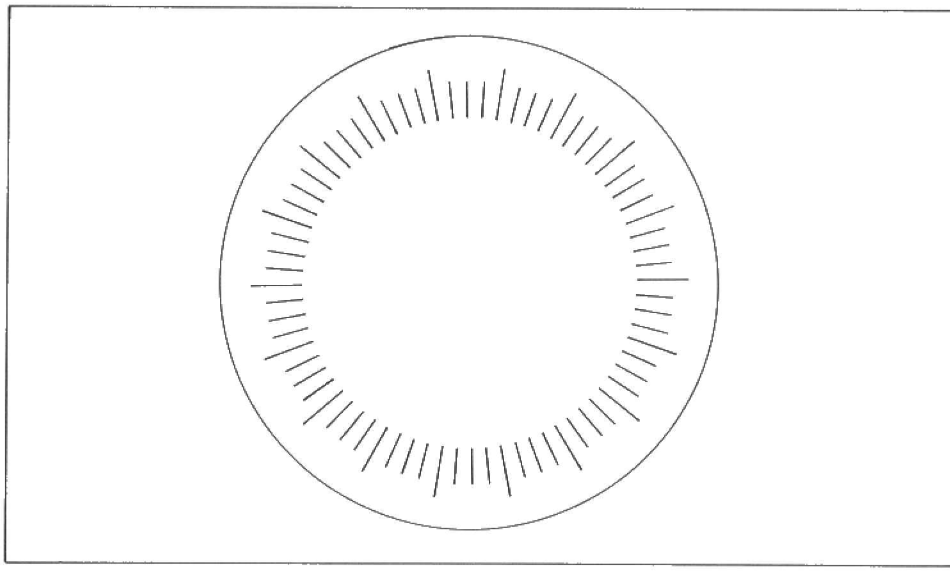


Fig. 7. Tuning vernier. This should be provided with a frequency scale after calibrating the receiver.

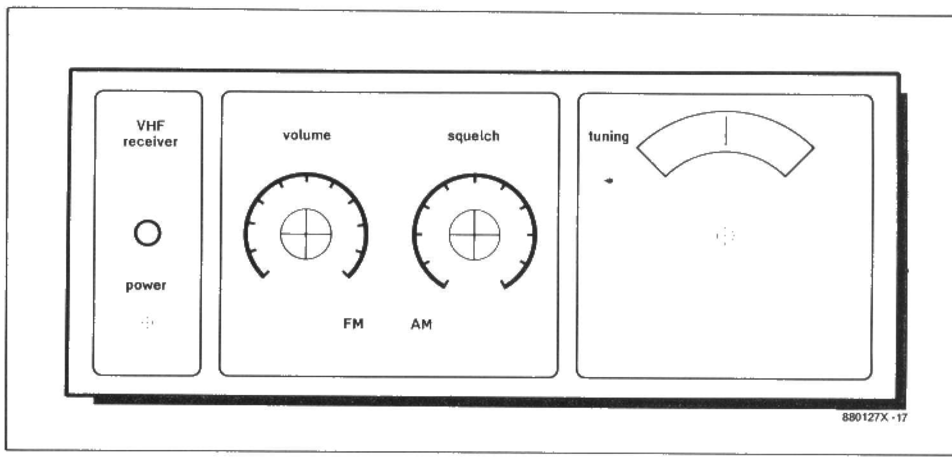


Fig. 8. Suggested front-panel layout.

NEWS

Top award for TV graphics and editing systems

Britain's top engineering prize, the 1988 MacRobert Award has gone to Quantel, creators of the Paintbox TV graphics system and the Harry video editing facility. The prestigious award is made annually by The Fellowship of Engineering, the UK's engineering academy, in conjunction with the MacRobert Trusts, to mark excellence of achievement in engineering innovation along with technical and commercial development.
Paintbox is a complete electronic graphic design system for programme production working directly in the TV medium. Using just a pressure-sensitive pen on a touch table, artists can effectively paint directly on to a video screen.
Harry is a video recording and editing system developed as a logical extension of Paintbox, which allows designers to work with the moving image. It is said to be the only system that can display the clips of video being worked on, in a similar way to that in which film is traditionally edited.

Amplifier for fibre optic systems

Avantek has introduced a 0.1 MHz to 4,000 MHz amplifier that offers 19 dB gain, ±0.5 dB full-band gain flatness, low pulse overshoot (less than 15%) and less than 1.8:1 input and output VSWR (all typical).
Designated ACT-4032, the amplifier is particularly suitable for use in high data rate (>1 GHz) fibre optic systems, as well as in pulse amplification, and instrumentation applications. The unit is also extremely versatile as a 'workbench' amplifier for the R&D laboratory.

New name for Teleprinter Group

At the recent AGM of the British Amateur Radio Teleprinter Group, members voted in favour of the proposed change of name to British Amateur Radio Teledata Group.

Brown goods market 1988

The figures for the third quarter of 1988 published by the British Radio & Electronic Equipment Manufacturers' Association suggest that:

- the colour TV market will achieve record levels again and complete the eleventh year of continuous growth in this market;
- teletext offtake is likely to achieve a year on year growth of 25% and will exceed the 1 million level for the first time;
- the FST market has continued to accelerate and, from 50% penetration at the end of 1987, will comprise some 80% of large-screen offtake for 1988.
- in the video recorder market, consumer activity has been very lively throughout 1988 and final figures should show an offtake of close to 2.4 million, substantially surpassing the previous record year of 1983.
- the audio sector has also been notable for the continued buoyancy of demand for compact disc facilities, particularly in the CD separates market which is likely to enjoy in excess of 30% growth.

PRACTICAL FILTER DESIGN (2)

by H. Baggott

Each filter has its own typical properties and these can be laid down in a few parameters. The second part of this series explains what these parameters are and what they mean

There are a number of parameters that characterize the properties of a filter. One of these is the frequency response characteristic or curve. The designer, having drawn up a target specification for the ripple in the pass-band and the slope of the filter skirt, will have to make a choice from several possibilities. The type of filter, whether it is a high-pass or band-pass, and so on, is not of importance at this stage.
Any type of filter can be converted into a standard low-pass with a cut-off frequency of 1 Hz. The target requirements must be translated into a normalized low-pass filter specification. After that, they may be compared with available standard curves with a 1 Hz cut-off point.
After a choice has been made, the required filter is simply reconverted and dimensioned for the required frequencies.
The designer has a choice of the following filter types:

- Butterworth
- Bessel
- Chebyshev
- transition
- linear-phase
- synchronous-tuned
- elliptic-function.

Apart from those of elliptic-function filters, the frequency characteristics of all these types are normalized for a −3 dB cut-off point at 1 Hz. The curves may be scaled to the desired frequency with the aid of standard multipliers.

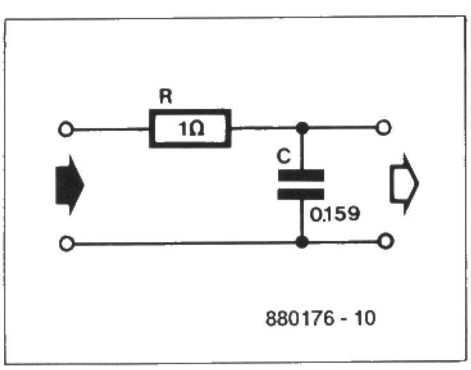


Fig. 6. A simple RC network with its −3 dB cut-off point at 1 Hz.

Filter parameters

As an example of the operation of a filter, we will consider the simplest type: an RC network as shown in Fig. 6. This network is terminated into an infinitely high impedance and powered by a voltage source that has an internal resistance of zero ohms. The capacitor is the frequency-dependent element and it introduces a phase shift.
The transfer function of the filter is

$$T(j\omega)=1/(1+j\omega CR) \qquad [4]$$

The absolute value of the function is

$$|\,T(j\omega)\,| \;=1/[\sqrt{1+(\omega CR)^2}] \qquad [5]$$

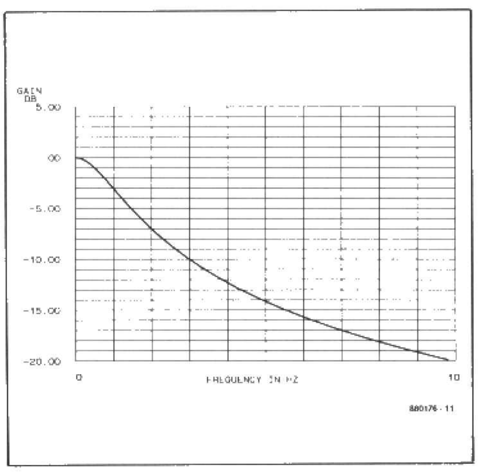


Fig. 7. Gain vs frequency curve of the sample filter drawn on a linear scale.

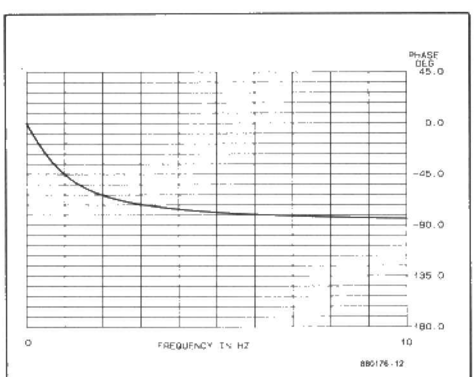


Fig. 8. The linear X-axis on the phase shift curve gives a good idea of the time delay caused by the filter.

The resulting phase shift is

$$\Phi=-\arctan(\omega RC) \qquad [6]$$

Equations [5] and [6] enable the gain vs frequency and the phase shift vs frequency characteristics to be computed and these are shown in Fig. 7 and Fig. 8 respectively. It should be noted that these curves are drawn on linear coordinates and that, therefore, particularly the gain curve is not the nearly straight line usually encountered. This is because the curves are normally drawn on a logarithmic abscissa (X-axis).
None the less, the phase characteristic in Fig. 8 shows how well the filter function approaches the condition not to

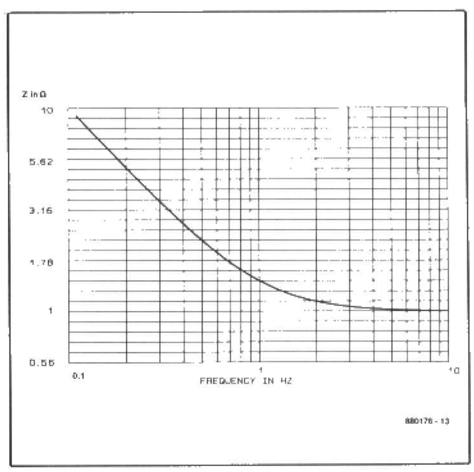


Fig. 9. The input impedance of the sample network is not constant but increases at low frequencies.

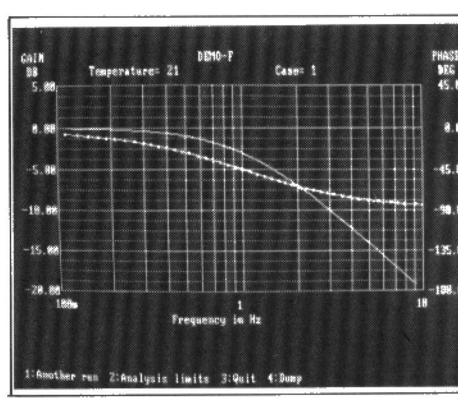


Fig. 10. The phase shift and gain characteristics are normally shown on the same illustration.

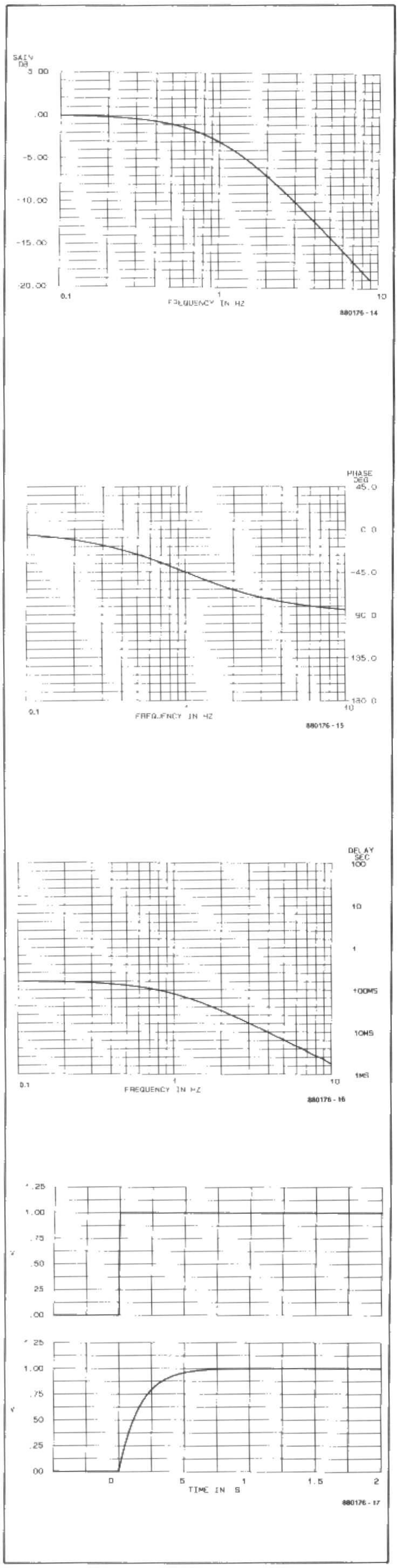


Fig. 11. The standard curves that will be used for all the filters that will be discussed in forthcoming articles in this series.
a. gain vs **frequency;**
b. phase shift vs **frequency;**
c. time delay vs **frequency;**
d. step response.

introduce delay distortion (Φ/f = constant). On a linear scale, the curve should be a sloping straight line. This aspect is difficult to judge when a logarithmic scale is used.

The input impedance of the filter is, of course, also a point to be considered. It is not possible, as many of us have found by bitter experience, to connect a number of filters in cascade to obtain a sharp cut-off response. Since the reactance of some filter components is frequency-dependent, the input impedance will also vary with frequency. This may be seen from Fig. 9, which illustrates the input impedance of our sample RC network.

Furthermore, a filter is always computed for a fixed ohmic termination. If that load is replaced by another filter presenting a frequency-dependent impedance, neither of the two filters will behave as originally designed.

As already stated, since the frequency and phase characteristics are normally drawn on a logarithmic X-axis (and quite often shown together as in Fig. 10), it is difficult to ascertain the time delay from them. For that reason, the time delay characteristic (computed from the frequency and phase characteristics) is often added on the same illustration.

For some applications, it is important to known the step response of the filter. This is a measure of the reaction of the network to a sudden rise in input voltage.

The four parameters just discussed give virtually all the information the designer normally requires.

Standard curves

The standard curves of our sample *RC* network are shown in Fig. 11. Such curves will also be given for all types of filter in forthcoming articles in this series. We will endeavour to give them all on the same scale so that a direct comparison may be made. All curves have been computed with the aid of a network analysis program to obtain representative characteristics that are as accurate as possible. All of them have been normalized on a cut-off frequency of 1 Hz.

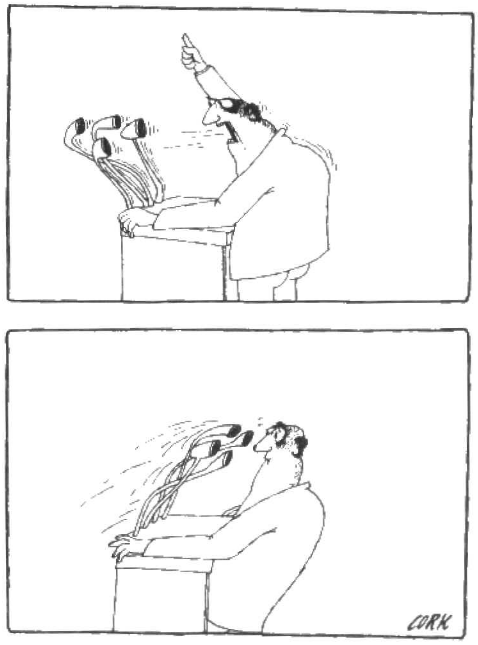
Reverting to Fig. 11, a and b show the gain vs frequency and the phase shift vs frequency respectively. Fig. 11c is the time delay vs frequency curve computed from curves a and b. Fig. 11d gives the step response of the network: the upper part of the illustration shows the sudden increase in input voltage from 0 to 1 V, and the lower part the resulting change in output voltage. Curves in future articles in this series will not show the upper part again, because the rise in input voltage is always taken as shown here. The step response of our sample filter does not mean much, of course, because the network is so simple. In the case of more complex networks (of the second and higher orders), the step response will show at a glance whether there is any ringing, how long this lasts, and the extent of the overshoot.

A sample computation

To end this second part of the series, we will give a sample computation to show how a filter is dimensioned in line with the foregoing discussion. Assume that we need an *RC* network as shown in Fig. 6 that is powered from a low-impedance voltage source and is terminated into a fairly high impedance(>1 MΩ). The cut-off point is required to be at 3 kHz (the multiplier, m, is thus 3,000). We choose a standard value for *R*, say 10 kΩ. The value of the capacitor is divided by the value of the resistor and the multiplier. If the network had contained an inductance instead of a capacitor, the value of the inductor would be multiplied by the value of the resistor and the result divided by the multiplier. In the *RC* network:

$$C = 0.159/mR =$$
$$C = 0.159/3000 \times 10000 =$$
$$C = 5.3 \times 10^{-9} = 5.3 \text{ nF}$$

The time delay at a given frequency may be calculated by reading the delay at that frequency in Fig. 11c and dividing that value by m. The same applies to the time scale of the step response curve.

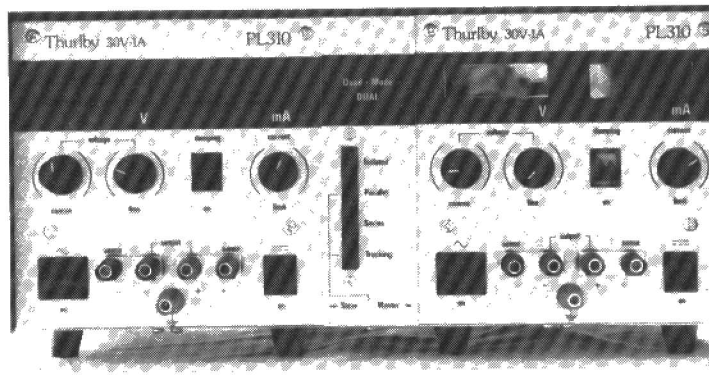


TEST & MEASURING EQUIPMENT

by Julian Nolan

Part 14: Power Supplies (2)

Thurlby PL310 QMD



Thurlby are a well-established Cambridgeshire based equipment manufacturer. Their innovative product range includes low-cost logic analysers and a microprocessor-controlled digital multimeter.

The PL310 QMD is a four-mode power supply (hence QMD) that has two main outputs and features four digital meters (3.75 digit) and current meter damping. It is part of a large range of power supplies manufactured by Thurlby that includes the low-cost LB-15 supply (4.5 – 15 V at 2 A) and the PLK series of triple output units. The PLK350K has the interesting feature of a 5 V/7 A (max) output.

Compared to other similar instruments, the PL310 QMD is relatively large, which is largely brought about by the space required to house the four 3.75 digit meters that monitor current and voltage on both outputs simultaneously.

Operating characteristics.

At 0.01% for a ±10% change in mains voltage, the line regulation should present no problems. The same may be said for the load regulation: <0.01% change in output for a 50% change in load. On the review model, both these figures were only about half these values.

The temperature coefficient of <0.01% per °C enables accurate output levels to be maintained, which is of particular value in view of the digital meters.

Separate output switches are provided to enable current and voltage adjustment to take place out of circuit.

Ripple and noise levels, typically below 1 mV, are good over the whole output range.

Maximum output voltage on the review model was 32.15 V, while the maximum current was 1.1 A.

Transient response time is good: the supply recovers to within 50 mV in 20 μs following a 100% change in load.

The output impedance is low (5 mΩ at 1 kHz) compared with some other units (typically 0.1 Ω) and this should enable the PL310 QMD to cope well with a wide variety of loads.

The PL310 QMD in use.

The PL310 QMD incorporates a versatile system of switching the four output modes. The ISOLATED mode enables both outputs to operate independently, while the PARALLEL and SERIES modes are self explanatory. The parallel mode permits outputs of up to 2 A (0 – 30 V), while in the series mode a voltage range of 0 – 60 V at up to 1 A is available.

While both serial and, normally, parallel operation may be improvised by external connections on power supplies that do not have these facilities as standard, the tracking mode can not. This mode enables twin matched outputs of ±30 V to be set conveniently by a single control. Furthermore, a continuously variable range of 0 V to 60 V can also be set by a single control.

It may be noted here that a twin supply, i.e., without the QMD option, is also available at £19 less than the QMD version. In my opinion, the extra £19 are well worth considering, because the four-mode facility makes the supply much easier to use when it is required to change between the various operating modes. Also, it has the added advantage of the tracking facility, which is likely to be especially useful in analogue work.

The 3.75 digit meters provide a clear and very accurate indication of the output parameters. They enable the PL310 QMD to be used in a wide range of applications where this sort of accuracy is required.

Typically, the output current can be set to an accuracy of ±2 mA, and the output voltage to an accuracy of ±10 mV.

Table 21

TECHNICAL SPECIFICATION

Line voltage: 110 V; 130 V; 220 V; 240 V ±10%
Line regulation: <0.01% of max. output for a ±10% change in mains voltage
Output impedance: <5 mΩ at 1 kHz
Temperature coefficient: 0.01% per °C (typical)
Transient response: <20 μs for output to recover within 50 mV following a 10% to 100% change in load
Ripple and noise: <1 mV (typical)
Voltage control: coarse and fine controls
Current control: single control only
Protection: against full overload and short circuits
Metering: 3.75 digit meter for current and voltage on each output
Meter accuracy: voltage 0.1%; current 0.3%
Output terminations: 4 mm terminals
Sensing: remote voltage sensing only
Other features: current meter damping; four operating modes, inlcuding tracking (accuracy ± 3% of setting); output switching
Dimensions: 175 × 345 × 255 mm (H × W × D)
Weight: 7.0 kg

The meter resolution is 1 mA.

Damping is a useful feature of the current meter and helps to overcome some of the inherent disadvantages of digital meters. It makes it, for instance, more usable in cases of a varying load.

Another impressive feature of the PL310 QMD is the facility to set the maximum

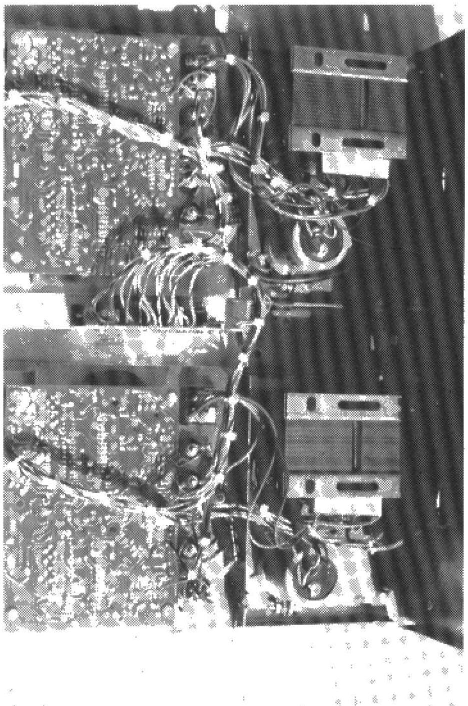
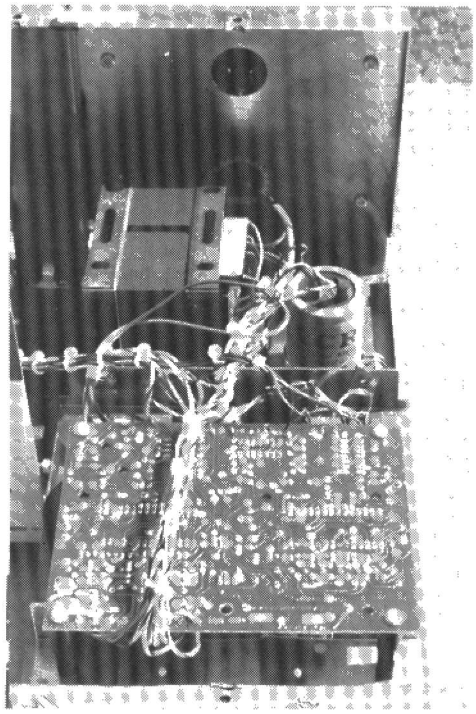
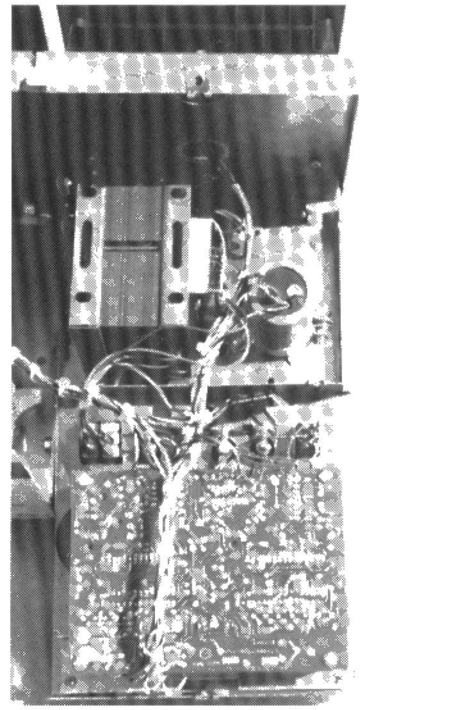




Table 22

	Unsatis-factory	Satis-factory	Good	Very good	Excellent
Voltage control					*
Current control					*
Regulation				*	
Meter accuracy					*
Overall accuracy				*	
Output impedance					*
Internal construction				*	
External construction			*		
Overall specification				*	
Ease of use				*	
Manual			*		
Additional features				*	

output current accurately before switching the supply into circuit. The maximum output current is then displayed, while fold-back current limiting is indicated by the flashing of the decimal places on the display.
Voltage sense terminals are provided that allow accurate voltage levels to be maintained regardless of the supply lead resistance or output current.

Construction.

The external construction is based on a stove-enamelled top cover on a steel chassis.
A neat front panel layout is provided by a plastic overlay with all the controls and their functions clearly marked.
The heat sinks at the rear of the instrument are covered.
In general, external construction is good, although it is noticeable that the two PL310 supplies and the QMD panel are essentially three separate units.
Internal construction is of a high standard. Each output is based on a single PCB that houses both the regulation and the DVM circuits, which should make servicing of the PL310 QMD an easy matter.
Heat dissipation is low with the obvious exception of the heat sinks when high currents are drawn.
In view of its rugged construction, the PL 310 QMD is, no doubt, capable of being used in a wide variety of operating conditions.

Manual.

The manual is intended for the entire range of the Thurlby PL series of power supplies and contains clear instructions for operating the supply in all its modes. Neither applications information nor circuit diagrams are included, but a service manual is available separately.

Conclusion.

Overall, the PL310 QMD represents a good compromise between price and performance. The quad mode facility should be of particular value to users who require ease of use, since it enables the versatility of the instrument to be enhanced without the necessity of external interconnection of outputs.
From most aspects, the construction is good and should enable the instrument to operate effectively in business, commercial and educational applications.

In perspective.

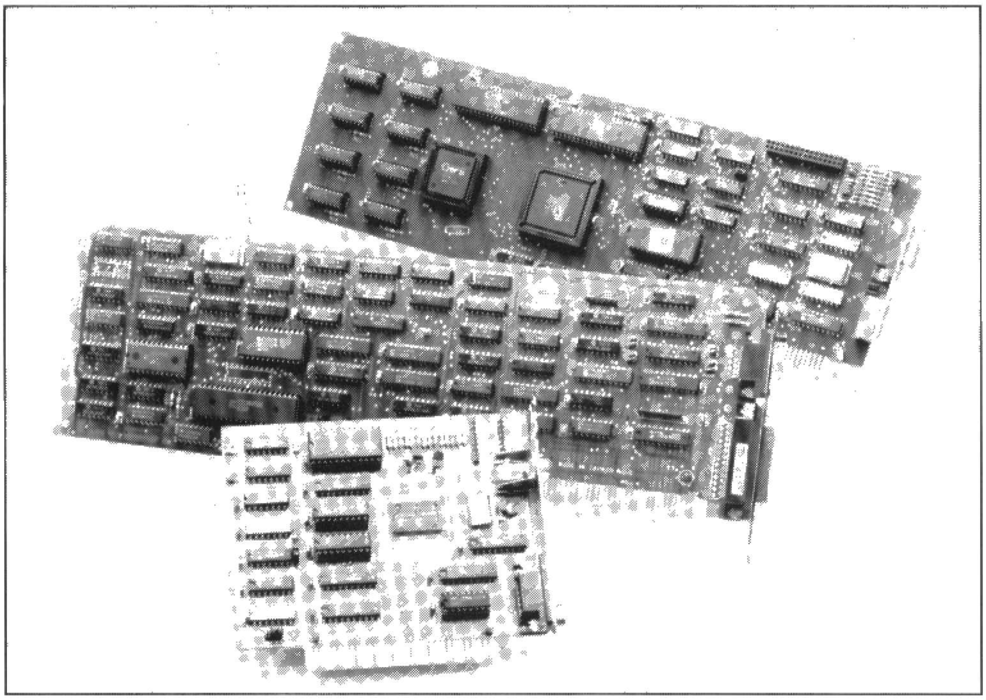
At £269 RRP the PL310 QMD compares very favourably with other power supplies available in this range.
It may be instructive to investigate the market for power supplies of around the £280 mark and offering a specification similar to that of the PL310 QMD, but with a 2 A maximum output (if this is required), such as the Thandar TS3022. It is, however, doubtful if any of such instruments offer the QMD facility.
The twin version mentioned in this review has an RRP of £250.

The PL310 QMD was kindly supplied by **Thurlby Electronics Ltd, New Road, St. Ives, HUNTINGDON PE17 4BG, Telephone (0480) 63570.**

VIDEO CARDS FOR PERSONAL COMPUTERS

by H. Stenhouse

In recent years a bewildering variety of video cards for PCs has come on to the market. This article attempts to remove much of the confusion caused by the different specifications and monitor requirements.



Functionally, the video card in a personal computer is an output device. Over the past few years, as PCs grew more sophisticated and users more demanding, the video card has become more than the fairly simple text display circuit of yesteryear. At that time, no provision was made for displaying, say, a graph on the screen. Fortunately, this was corrected with the introduction of the Colour Graphics Adaptor (CGA), which did allow, at least partly, for integration of text with simple graphics. The main disadvantage of the CGA was, however, its limited resolution for text. The well-known Hercules card, developed by the company of the same name, overcame this shortcoming at least for monochrome text applications. A few years later, the EGA card (EGA = Enhanced Graphics Adaptor) and the PGC card (Professional Graphics Adaptor) were introduced to satisfy more demanding users wishing to work with high-resolution colour screens. But the evolution of the videocard did not stop with the PGC: the introduction of the new series of PS/2 computers from IBM called for even higher resolution and speed: the answer was provided in the form of a range of MCGA and VGA cards.

The evolution from the basic video card to the highly sophisticated graphics adaptor available now has caused great confusion among many PC users. This is mainly because the systems are often incompatible as far as the monitor, horizontal and vertical scanning frequency, and even the interconnecting cables are concerned. The CGA (8 colours) and the EGA card (16 colours), for instance, supply output signals at TTL level, usually combined with an intensity signal, whereas other videocards, such as the PGC and VGA have linear video outputs that allow a very high number of displayable colours. Owing to the structure of the on-board RAM, the VGA and PGC work with video modes in which a limited number of colours — say, 256 — can be active at a time. These cards offer an indirect choice of nearly a quarter of a million shades via a palette structure.

Computer display manufacturers have traditionally supported each new PC video card with an appropriate display. An exception to this is formed by the so-called *multisync* monitor, which is available in many types from, for instance, NEC (Multisync-2), Eizo (Flexscan 8060S and 9070S) and Taxan. The electronics in this advanced type of display is capable of automatic adjustment to the internal line and raster frequency detected in the applied video signal. In addition to this extremely useful feature, the display often has inputs for linear as well as digital video signals, so that it can be used with virtually all current videocards.

Unfortunately, a CVBS (Chroma, Video, Blanking and Sync) input to the PAL (or NTSC) standard is rarely found on high-resolution colour monitors for computers. Such an input is, admittedly, not very useful in the PC environment, but may, on the other hand, give interesting opportunities for use of the high-resolution display in conjunction with cameras, VCRs, video digitizers, and some types of home computer.

The monochrome scene

The Monochrome Display Adaptor (MDA) fitted in the earliest of IBM PCs provided only text display. This card, which is now obsolete, had a screen memory of only 4 Kbyte (4,000 characters), and displayed text as 25 lines of 80 characters. None the less, the resolution of the MDA is relatively high at

720×350 pixels. The character font is 7×11 pixels in a 9×14 raster, resulting in a clear text display. The card provides 256 characters, which are stored in an on-board ROM. No provision is made for the user to define his own characters. The Hercules card replaces the MDA, and adds a graphics option in the form of a monochrome graphics interface with a resolution of 720×348 pixels. Text display is basically the same as with the MDA, and requires no special software. Graphics software, however, can only be run with the aid of software utility INT10, since IBM, and, therefore, the Disk Operating System (DOS), does not support the Hercules card. After an initial shortage of graphics software for the Hercules card, this is now supported in the majority of programs from leading software companies. The use of the Hercules card is also boosted by programs such as MG-2 (MultiGraph-2) that allow it to emulate the CGA mode with the aid of grey shades. Currently, the Hercules card is probably the widest used video adaptor for PCs running word-processing and other text applications. As a useful boon, the card provides a parallel printer output port, LPT1:.

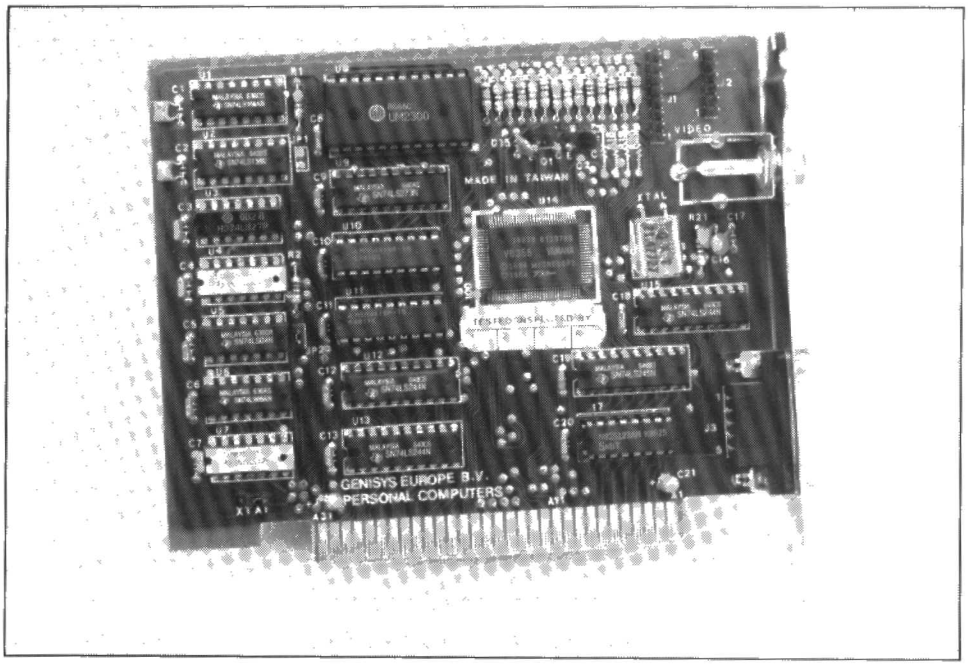


Fig. 1. Half-length Colour Graphics Adaptor.

Colour and graphics: the CGA

The CGA was the first card introduced by IBM that allowed the connection of a colour display to a PC. On board the CGA is a 16 Kbyte memory. The card can operate in two modes: text and graphics. In text mode, two sub-modes are available: 40 or 80 characters per line, at 25 lines per screen in both cases. The available memory allows 8 or 4 screens to be stored in 40 and 80 character mode, respectively, so that fast scrolling can be achieved. The graphics mode also affords a number of sub-modes, including one with 640×200 pixels at two colours, and 320×200 pixels at four colours. In graphics mode, characters with an ASCII value greater than 127 can be shaped by the user. Since characters are formed in an 8×8 matrix, the CGA is less suitable for text display. In many cases, a CGA and a Hercules card can be used alongside in the computer, but only if the Hercules card is not used in the so-called *full-size mode* (64 Kbytes of screen memory). Switching between the two cards can be done in software, so that monochrome text can be combined with colour graphics on separate screens.

Table 1.

Horizontal frequency kHz	Vertical frequency Hz	display adapter	Resolution
15.75	60	CGA	320 × 200 640 × 200
18.2	50	EGA-mono	640 × 350
18.4	50	MDA	720 × 350
18.8	51	Hercules	720 × 348
21.9	60	EGA	640 × 350 1056 × 352
31	60	EGA+	640 × 480
32	60	CGA double frequency	320 × 400 640 × 400

The CGA double-scan card is an improved version of the standard CGA. This type of video adaptor is available in the form of an emulated mode on some EGA cards, and enables software written for the CGA to be run on a display with much higher resolution. This is mainly by virtue of the double-scan principle, which provides an interlace function that effectively doubles the vertical resolution. Unfortunately, this interesting mode is not available in the form of a separate card.

Enter the EGA

The cost of an Enhanced Graphics Adaptor (EGA) was, for a time, prohibitive for the average PC user, but that, fortunately, changed with the availability of good-quality products from the Far East. The EGA has a large, 256 Kbyte, on-board memory, and offers a graphics resolution of 640×350 pixels, at 16 possible colours per pixel (a 256-colour extension for the EGA is described in Ref. 1). Pixel colour selection is from a 64-colour palette. Depending on the resolution, two or four screens can be held in the memory.

The character set of the EGA is ROM-resident, and uses an 8×14 matrix to guarantee excellent text display capabilities. Provision has been made for the user to shape up to 1024 characters at a height of 8 to 32 pixels. Many manufacturers of EGA cards have come up with useful extensions to the basic capabilities, often in the form of emulation modes. In many cases, software is supplied with the card that allows it to switch to the CGA, MDA and CGA double-scan mode. The EGA-Wonder card from ATI takes compatibility even further by its ability to adapt the outputs to the display used. Other EGA cards

Table 2

pin	MDA	CGA	EGA	PGC	MCGA	VGA mono	VGA colour
1	ground	ground	ground	R	R-out	-	R-out
2	ground	ground	R'	G	G-out	M-out	G-out
3	-	R	R	B	B-out	-	B-out
4	-	G	G	Com. Sync	-	-	-
5	-	B	B	mode	-	test	test
6	intensity	intensity	G'	R-ground	R-In	key	R-In
7	video	-	B'	G-ground	G-In	M-In	G-In
8	HSYNC(+)	HSYNC(+)	HSYNC(+)	B-ground	B-In	-	B-In
9	VSYNC(−)	VSYNC(−)	VSYNC(−)	ground	key	-	-
10					ground	G (dig)	G (dig)
11					type 0	-	G (dig)
12					type 1	G (dig)	-
13					HSYNC	HSYNC	HSYNC
14					VSYNC	VSYNC	VSYNC
15					-	-	-

can be used in conjunction with a CGA-compatible monitor, with the obvious advantage of going round investing in a new, high-resolution, monitor.

PGC: professional at a professional price

The Professional Graphics Adaptor (PGC) was developed and introduced to convince PC users of the fact that CAD software need not necessarily be run on a professional workstation. Unfortunately, the PGC has remained relatively expensive, and has, therefore, failed to become popular. Aimed at the CAD market, the PGC was designed to provide an aspect ratio of 4:3, and to generate up to 256 colours. The multisync monitor mentioned earlier makes it possible to run PGC-based CAD software in the EGA+ mode available on the latest multi-mode EGA cards. That the PGC is bound to be forgotten soon is also caused by the fact that the Video Grapics Array (VGA), introduced with IBM's line of PS/2 computers, is in principle capable of taking over all its functions... as a subset!

New standards: MCGA and VGA

In an attempt to put an end to the widespread confusion about videocards in PCs, IBM recently introduced two new types of display adaptor, the Video Graphics Array (VGA) and the Multi-Color Graphics Array (MCGA), for use in their Series PS/2 computers. Both adaptors are complete, versatile, and expected to stay with us for quite some time. The MCGA is essentially a 'low-budget' version of the VGA. It comes as standard with the Model 30 computer in IBM's PS/2 line, and has 64 Kbyte of on-board RAM. The maximum resolution of 640×480 pixels is achieved in the two-colour mode. At the lower resolution of 300×200 pixels, 256 colours are available from a total of 262,144 in the colour palette. The MCGA can display up to 64 grey shades on a monochrome monitor. Downwards compatibility is ensured at least partly by a CGA emulation mode. Other cards such as the EGA or MDA can not be emulated.

The second new card, the VGA, is fitted in PS/2 Models 50, 60 and 80. It can be used with colour as well as monochrome monitors with an analogue (linear) input. Screen memory is 256 kByte for a standard graphics resolution of 640×480 pixels, or 720×400 pixels in the text modes. In the low resolution graphics mode, each of the 320×200 pixels can be assigned one of 256 colours. In the high-resolution mode, this is reduced to 16 colours. The number of available colours in the palette circuit is equal to that in the MCGA. Depending on the selected screen mode, up to 8 screens can be held in memory. Characters are built in a matrix of 9×16 pixels in text mode, or 8×16 pixels in graphics mode. The VGA is capable of emulating all previous standards, ensuring software compatibility with MDA, CGA, EGA and MCGA. Characters in these subsets have a maximum height of 32 pixels.

Wanted: monitors!

Selecting a videocard is one thing, finding a suitable monitor for it is another. Table 1 summarizes the vertical and horizontal scanning frequencies of a number of PC videocards. In general, the requirements of the cathode ray tube (CRT) used in the monitor rise with sync frequency. Excellent resolution on a flicker-free display is achieved thanks to high raster and line frequencies (up to 80 Hz and 50 kHz respectively), and non-glare screens.

Obviously, investing in an expensive videocard is useless if the monitor has insufficient resolution. In the case of the monochrome monitor, the maximum resolution is usually determined by the bandwidth of the video amplifier. This means that the design and production of a monochrome monitor are simple compared with those of a colour monitor with equal specifications in respect of resolution. For optimum convergence in a colour picture tube, the electron beams must be controlled with great accuracy to ensure the actuation of only one phospor element at a time. The size of these elements varies from 0.62 mm in a standard colour TV tube to 0.29 mm in a multisync high-resolution colour monitor. For most graphics applications, a dot pitch of 0.31 mm is sufficient.

The large differences in respect of line and raster frequency between the various videocards give rise to monitor incompatibility. A standard CGA display, for instance, can not be used in conjunction with a Hercules card. Some monochrome, Hercules compatible, monitors, however, are capable of dual-frequency operation so that CGA pic-

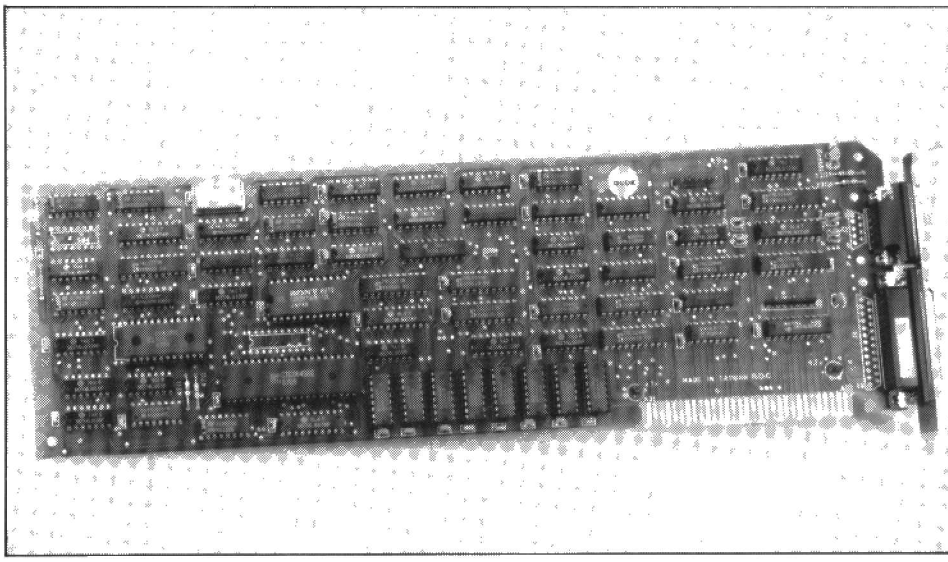


Fig. 2. Hercules card for combined medium-resolution monochrome graphics and text applications. The card shown here is a relatively old, full-length model.

Table 3. Videocard	Resolution	Colours
Hercules	720 × 348	2
CGA	320 × 200	2
CGA	320 × 200	4
CGA	640 × 200	2
EGA	320 × 200	16
EGA	640 × 200	4
EGA	640 × 200	16
EGA	640 × 350	2
EGA	640 × 350	4
EGA	640 × 350	16
VGA	320 × 200	256
VGA	640 × 400	16
VGA	640 × 480	2
VGA	640 × 480	16

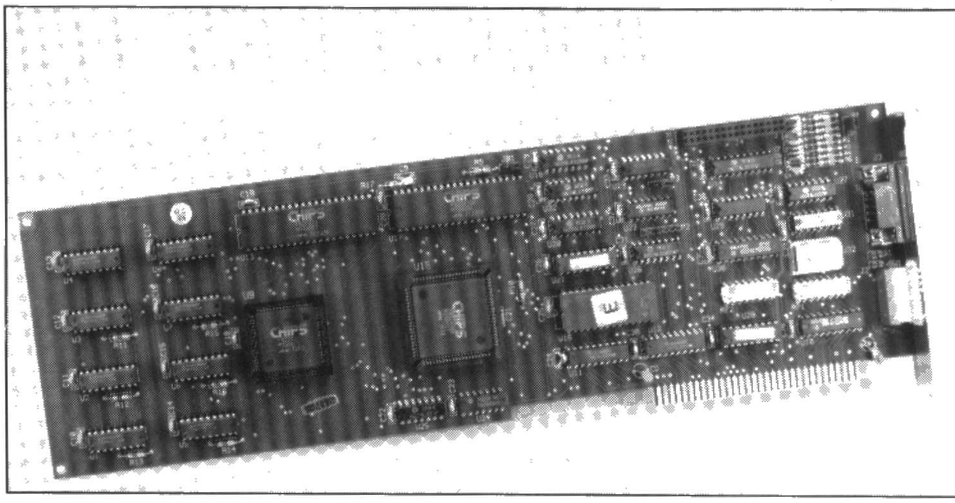


Fig. 3. The widely used EGA card affords good colour graphics and text capabilities at a reasonable price.

tures can be displayed by means of shades of grey. Similar dual-sync colour monitors are aimed at users of PCs with a CGA and/or EGA card. The EGA and PGC also require their own monitor type. Apart from a specific line and raster frequency, some videocards supply only digital or analogue signals. Potential buyers of a video card are, therefore, well advised to take all these different specifications into account before deciding on a particular type. Always remember the monitor!

A great effort is constantly being made by monitor manufacturers to provide the widest possible range of products to meet the requirements of customers as well as of the videocards they use. Again, the multisync monitor (monochrome as well as colour) is the overall winner here, although VGA and MCGA compatibility is not always guaranteed.

Cables and plugs

Combining a videocard with an appropriate colour or monochrome monitor is a problem that is even further complicated by the cables and plugs needed for each combination. Table 2 shows an overview of connections for various type of video card. It will be noted that the MDA, CGA, EGA and PGC make use of a 9-pin D-connector, while the new cards, MCGA and VGA, need more wires and work with a 15-pin connector. In priciple, horizontal and vertical sync signals are sent over separate wires; only the PGC uses a combined sync line. Fortunately, this card is of no significance to today's PC user.

Reference:

1. A 256-colour adaptor for the EGA. *Elektor Electronics* March 1988.

TOUCH-KEY ORGAN

A simple-to-build, inexpensive, two-octave organ with touch-sensitive keys etched on the printed circuit board.

by T. Wigmore

In spite of the fact that the present organ is monophone, and does not provide semitones, it is, none the less, a perfect musical instruction aid.

In the circuit diagram shown in Fig. 1, the central oscillator of the mini organ is formed by IC_5, the well-known timer Type 555. The output frequency of the 555 in astable multivibrator configuration depends on the values used in the R-C network connected between pins 6 and 7. In the present application, the output frequency of the organ is, therefore, determined by C_3 and the binary code applied to inputs A to D of 16-channel analogue multiplexer IC_2. This CMOS IC acts as a large, single-pole, rotary switch that connects one of resistors R_{12} to R_{26} incl. between R_{11} and R_9. If, for example, the one but highest tone is to be generated, the one but lowest resistor value must be selected from the ladder network. This means that output 1 (pin

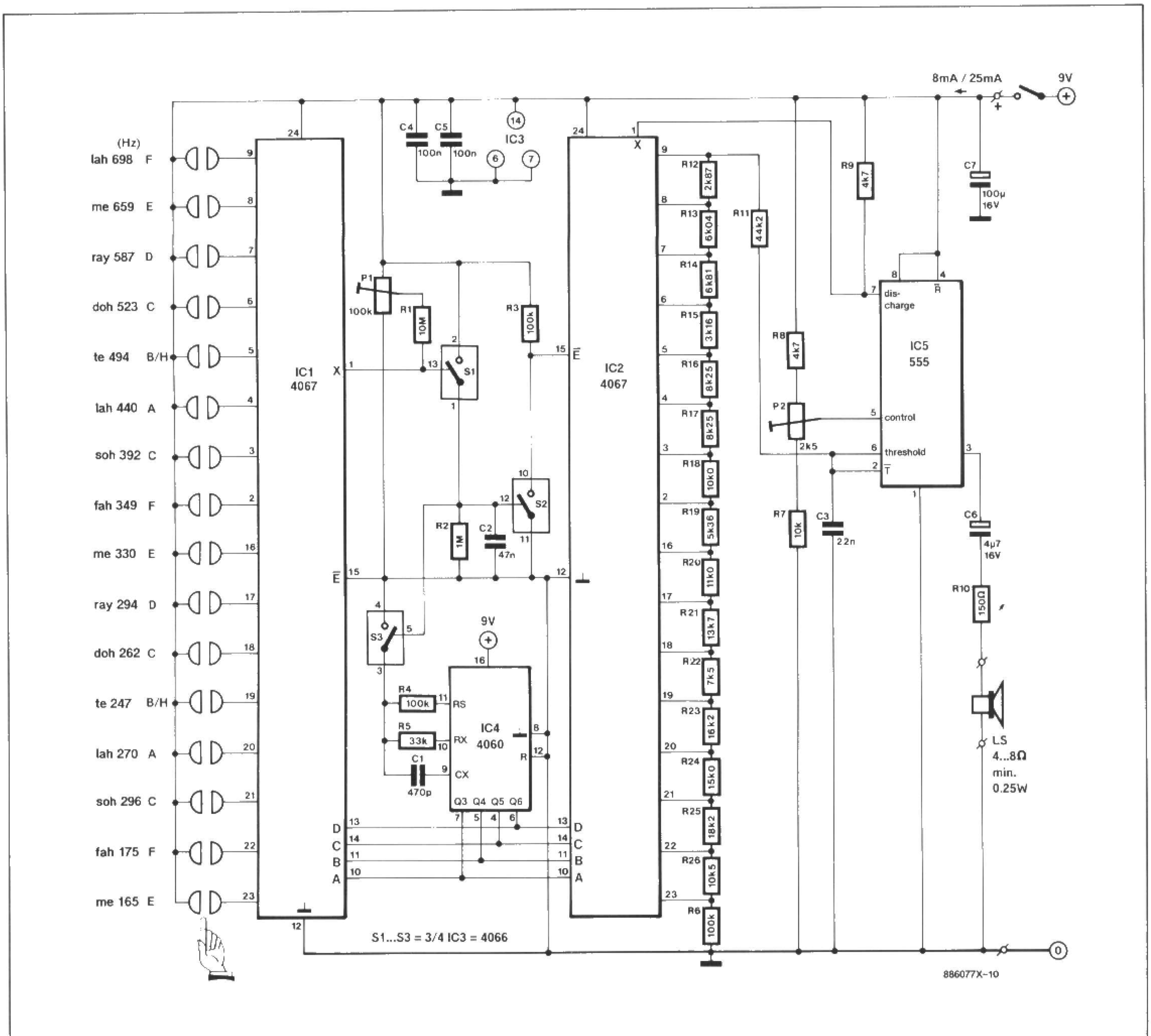


Fig. 1. Circuit diagram of the miniature touch-key organ.

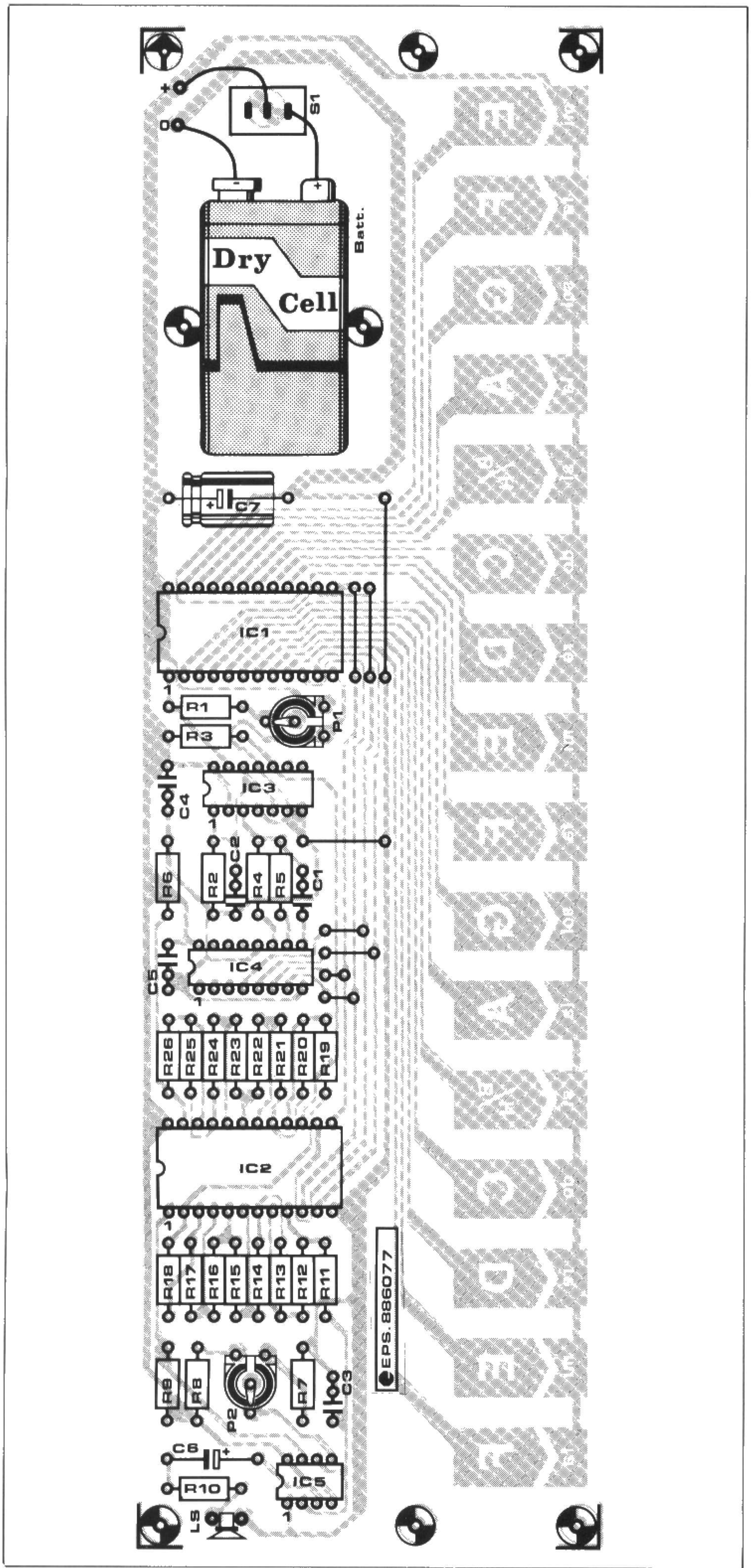


Fig. 2. Track layout and component mounting plan.

8) of IC_2 must be connected to the centre contact, X, at pin 1. This selection is achieved by applying binary code 1000 to the ABCD inputs: only bit A (pin 10) is logic high. Non-selected multiplexer outputs are at high-impedance, and the stated BCD code causes the 555 to oscillate at a frequency determined by network $(R_9+R_{11}+R_{12})$-C_3. The resistors in the ladder network are 1% types from the E96 series, dimensioned for the tone scale available on the mini organ.

The 16-key touch-sensitive keyboard is based on another CMOS multiplexer type 4067, IC_1. By virtue of the high input impedance of the CMOS bilateral switches in this chip, skin resistance between the positive supply line and the input selected by IC_4 can be detected. The address generator, IC_4, provides a key-scan rate of about 60 kHz. When a key is a touched, the potential at the control

Parts list

Resistors:
R_1 = 10M
R_2 = 1M0
R_3;R_4;R_6 = 100K
R_5 = 33K
R_7 = 10K
R_8;R_9 = 4K7
R_{10} = 150R
R_{11} = 44K2F
R_{12} = 2K87F
R_{13} = 6K04F
R_{14} = 6K81F
R_{15} = 3K16F
R_{16};R_{17} = 8K25F
R_{18} = 10K0F
R_{19} = 5K36F
R_{20} = 11K0F
R_{21} = 13K7F
R_{22} = 7K5F
R_{23} = 16K2F
R_{24} = 15KF
R_{25} = 18K2F
R_{26} = 10K5F
P_1 = 100K preset H
P_2 = 2K5 preset H

Capacitors:
C_1 = 470p
C_2 = 47n
C_3 = 22n; MKT
C_4;C_5 = 100n
C_6 = 4μ7; 16 V; axial
C_7 = 100μ; 16 V; axial

Semiconductors:
IC_1;IC_2 = 4067
IC_3 = 4066
IC_4 = 4060
IC_5 = 555

Miscellaneous:
S_1 = miniature on/off switch.
Ls = loudspeaker; 4-8 Ω; 0.25 W.
PCB Type 886077 (see Readers Services page).

input of S_1 rises to a level high enough to cause this electronic switch to be closed. If S_1 is closed, S_2 and S_3 will be closed also. Since S_2 then pulls the $\overline{E}$ (enable) input of IC_2 low, this chip actuates the internal bilateral switch selected by means of the applied BCD code. The keyboard scan oscillator is disabled via S_3 to ensure that the selected note sounds as long as the key is touched. Components R_2 and C_2 form a basic retriggerable monostable that serves as a key debounce circuit.

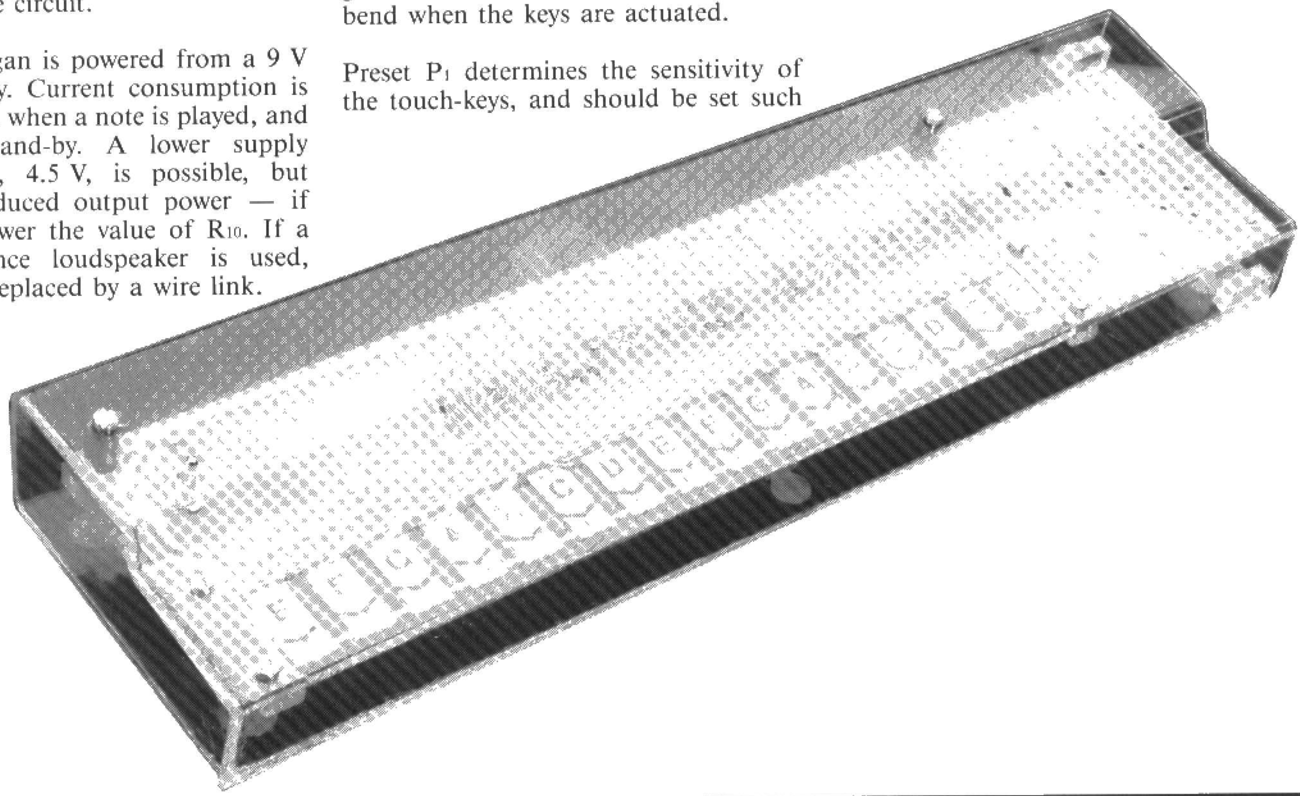
The mini organ is powered from a 9 V (PP3) battery. Current consumption is about 25 mA when a note is played, and 8 mA in stand-by. A lower supply voltage, e.g., 4.5 V, is possible, but results in reduced output power — if necessary, lower the value of R_{10}. If a high-impedance loudspeaker is used, R_{10} may be replaced by a wire link.

Construction and alignment

The construction of the mini organ on ready-made, single-sided, printed circuit board 886077 is a matter of routine, although care should be taken not to overlook any of the wire links on the board. The PCB supplied through the Readers Services is pretinned to prevent oxidation of the key contacts. The photographs in this article show how the organ is built from pieces of perspex to guarantee a compact unit that does not bend when the keys are actuated.

Preset P_1 determines the sensitivity of the touch-keys, and should be set such that the note generator is just off when none of the keys is touched (key threshold level).

The mini organ is tuned with the aid of P_2, which is adjusted for an output frequency of 440 Hz when reference note A (*lah*, 11th key from the left) is played. This adjustment can be carried out either with a frequency meter or a tuning fork.



NEWS

Aero SATCOM 'first' for Racal

Racal Avionics is to supply the world's first production aeronautical satellite voice and data system under a major contract from the US Gulfstream Aerospace Corporation.

Racal will provide three single-channel voice and data satellite communication systems for installation in the Gulfstream IV corporate business jet aircraft. Gulfstream has options for the purchase of at least seven more systems within two years.

Advanced 200,000 gate CMOS cell-based technology from LSI Logic

LSI Logic has recently unveiled the industry's first cell-based custom ASIC technology capable of integrating 200,000 equivalent logic gates on a single chip, enabling customers to achieve the complexity of two-and-a-half VAX*780 minicomputers on a single ASIC. In addition to its very high gate capacity, a key feature of the LCB007 is the amount of high-density memory that can be designed into a single chip. Fast static RAM and ROM complexities of up to 144 kbits and 1 Mbit, respectively, are possible.

*VAX is a trademark of Digital Equipment Corporation.

World's first telepoint demonstration

The world's first public demonstration of the revolutionary Phonepoint mobile phone service was carried out by British Telecom recently from a BT Phonepoint in Euston Square, London.

Phonepoint is BT's version of the public telepoint service being pioneered in Britain and based on a new generation of cordless phones known as CT2. It will allow users to make calls from a network of Phonepoints situated in public places — bus and railway stations, shopping centres, airports, garages, and motorway service stations.

The Phonepoint service will be underwritten by guarantees of quality and reliability, with refunds to customers if the specified performance is not met. Phonepoint will employ independent auditors to check reliability and quality and will publish the results.

AEG-Siliconix partnership

AEG AG and Siliconix Inc. have recently announced their agreement to work together on a range of activities in the field of power MOS and smart-power semiconductor products under long-term co-operation and licenses. AEG will also purchase 39% of Siliconix common shares.

The companies have agreed that Siliconix will grant AEG certain patent licenses for the design and manufacture of semiconductor products. They have also agreed to share know-how, to exchange the results of research, and to combine their engineering efforts for the development of certain new power MOS applications and products.

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TWEETER PROTECTOR

by K. Baumotte

Tweeters, the high-frequency drive units in a loudspeaker system are often damaged by a properly matched and rated power amplifier. This happens because many modern power amplifiers are direct-coupled to the loudspeakers, i.e., they do not use a transformer. Such amplifiers have the nasty property of producing square-wave signals when they are (even slightly) overdriven. The ensuing harmonics lie chiefly in the frequency range of the tweeter. This considerable spectral shift of the audio signal was, of course, not taken into account during the design stages. This is because during the standard (DIN) testing of the loudspeaker system, the tweeter is required to handle only 1% of the total applied power. In other words, when 100 W of audio power is applied, to a loudspeaker system, the tweeter needs to handle only 1 W. Even if the tweeter is rated well above the standard test specification requirements, during loud music passages, when clipping occurs (and square-wave signals are generated) it may well have to handle too much power. This may happen before any distortion of the sound is heard. Tone controls and equalizers can hasten the demise of the tweeter: a 6-dB lift at 4 kHz doubles the power applied to the tweeter, i.e., makes the unit twice as vulnerable.

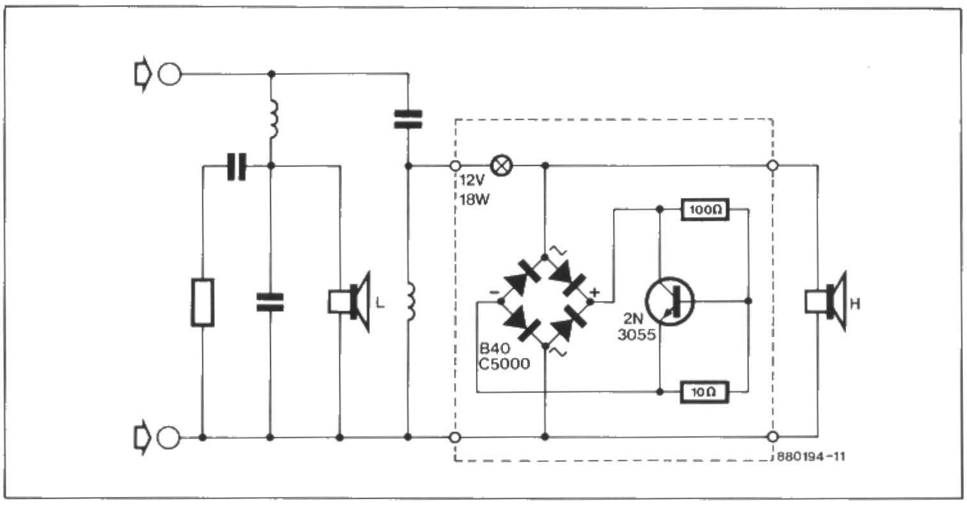


Fig. 1. Circuit diagram of the proposed tweeter protector. The resistance of the lamp increases with rising input power. At the same time, the 2N3055 short-circuits the tweeter during the high power peaks. The circuit is suitable for all tweeter with a cross-over frequency of 5 kHz.

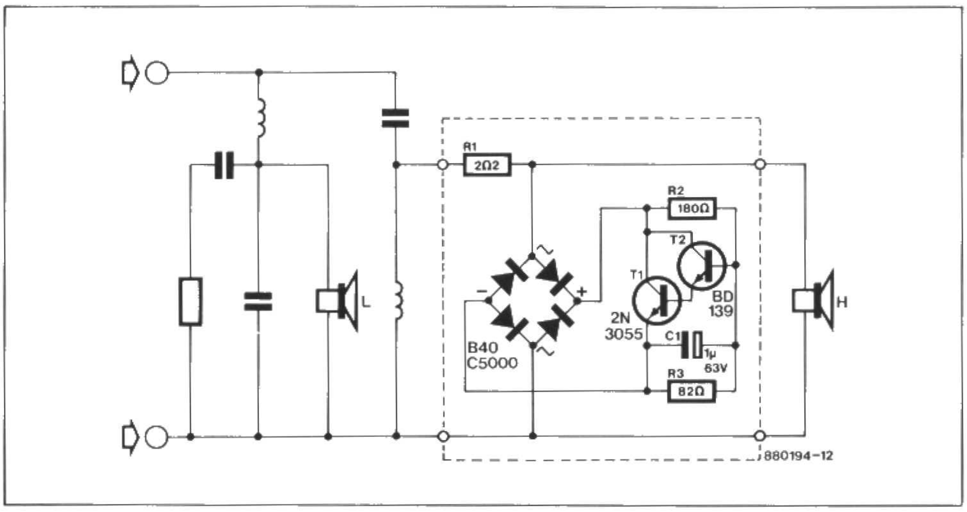
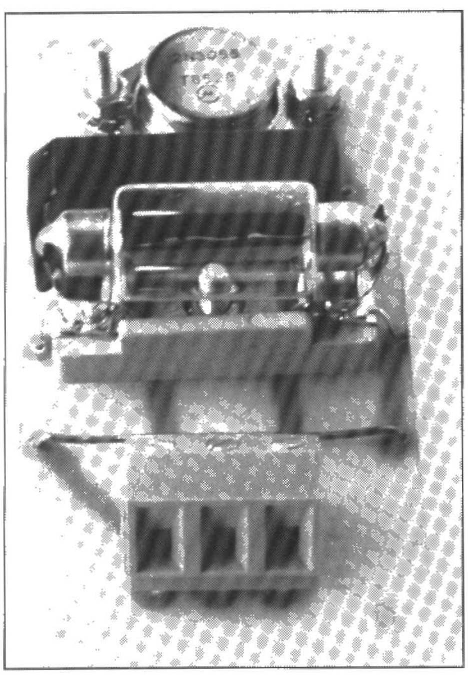


Fig. 2. This is an improved version of the circuit of Fig. 1. Its starting time and time constant may be varied to individual taste. Moreover, it keeps distortion down to an acceptable level.

Protection circuit

A relatively simple circuit as shown in Fig. 1 is all that is required to prevent damage to the tweeter, especially if it is frequently used at high volume levels. It can not be a coincidence that most leading suppliers of disco and public-address systems fit a similar protection circuit in their equipment.



A useful side effect of the circuit in Fig. 1 is that the lamp used as a positive-temperature coefficient resistor gives a visible warning if the sound level is too high. The lamp begins to glow when the power reaches a certain level. When the voltage drop across it reaches about 5.5 V, the lamp severely limits the level of the applied signal. At the same time, the transistor begins to conduct and short-circuits the tweeter. If this situation is allowed to continue, the transistor dissipates enough heat to warrant the use of a small heat sink.

Distortion

Overloading the tweeter also causes severe distortion: when the voltage at the input to the circuit is 12 V, the level of distortion is likely to be around 10%. This may be improved very considerably by the use of the circuit shown in Fig. 2. Under the same conditions, the level of distortion is only about 0.2%.

The time constant R3-C1 enables single pulses to pass through the circuit unhindered. Only when the overloading continues do the darlington pair of transistors begin to conduct and short-circuit the drive unit. The time constant may be altered by changing the value of C1 up to 470 µF. Slight alterations in the values of R2 and R3 allow the toggle time of T2 to be set to individual needs.

Finally

Although the circuits in Fig. 1 and Fig. 2 are extremely useful, it should be noted that they are not intended for use with good hi-fi equipment: they are designed primarily for use with disco and public-address equipment.

Evening on the planet Oldana, as the Maplin Juggertrain thunders along the highway; captured on canvas by galaxy famous artist Lionel Jeans and featured on the cover of the new Maplin Catalogue.